IBA

TECHNICAL REVIEW

12

Techniques for Digital Television

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Contents

Introduction

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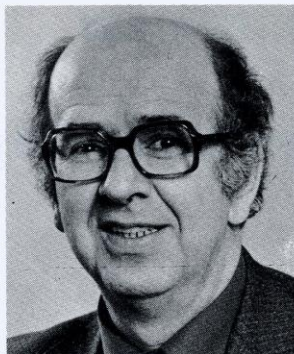
As with certain other branches of human endeavour the development of a technology can be likened to the life of a climbing rose-tree. In the first place the underlying principle of the particular technology is budded on to the root stock of scientific understanding, growing in the fertile soil of man's imagination and nourished by the industrial climate prevailing at the time. With careful management, and not without an element of good luck, the bud will 'take' and the briar become a rose. At this point it might happen that some of the less promising ones must be grubbed out in order to enhance the effect of natural selection, and this can be just as heartbreaking in the analogy as in the event that it represents. In good time, it is hoped, the efforts of all concerned will be rewarded as some of the shoots bear flower.

The tree, thus established, may continue to bloom, slowly proceeding from strength to strength. Occasionally as a result of a mutation a much more virile new stem, a 'sport', may appear bursting out from the base, so strong and vigorous, that it channels all further growth through itself and eventually causes much of the older wood to wither and fade.

Such a thing is now taking place within the technology of broadcasting. The new 'sport' that is channelling most of the present and likely future development in this sphere represents digital techniques, and already some of the gnarled, older branches of analogue wood, that have done so well in past years, are being threatened by the pruning knife.

Work on some of the applications of digital techniques to broadcasting has already been featured in three previous issues of *IBA Technical Review*, but so important are these changes that we need make no apology for this further volume. It contains descriptions of a number of items of equipment designed and constructed by the IBA which together formed the essential elements of a digital television studio based on so-called sub-Nyquist sampling. These, together with a number of contributions from other organisations, were jointly demonstrated to the EBU Technical Committee in Venice during April 1977 with a view to submitting proposals to the CCIR for the adoption of agreed digital standards.

This healthy young 'sport' of digital video processing, which has already flowered so successfully in specific fields, promises a prolific display of blooms in due course.

An Introduction to Sub-Nyquist Sampling

by K H Barratt and K Lucas
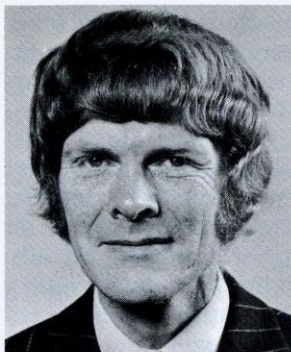
KENNETH BARRATT, BSc, CEng, MIERE, joined the Authority in 1970 and worked initially on the use of digital computers for monitoring uhf television transmitters. In 1973 he joined the Video and Colour Section as manager of the DICE project. A graduate of Reading University his previous experience included four years with BBC Designs Department working on the design of uhf television transposers.
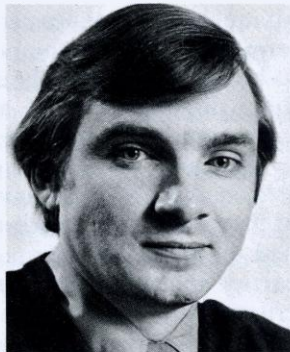
KEITH LUCAS, MSc, PhD, graduated in 1967, and afterwards spent four years at Southampton University on research in the field of Artificial Intelligence and Adaptive Control Systems. He was then employed by the Plessey Company and worked on a number of defence projects. In 1974 he joined the Authority's Experimental and Development Department where, within the Automation and Control Section, he was engaged in the development of ORACLE. He became Head of Video and Colour Section in the early part of 1978.

Synopsis

At the present time the technology of digital television is being actively studied in Europe, North America and Japan, and it seems likely that digital techniques will eventually predominate in the studio. For this reason members of the EBU have been devoting their attention to the consideration of future standards for digital television. While a number of countries favour a multiplexed YUV component format, the UK has expressed a desire to adopt a standard based on composite PAL and foresees the use, wherever possible, of a composite PAL signal sampled at the sub-Nyquist rate of twice sub-carrier frequency.

This article explores the reasons for this choice and presents the theoretical basis of the sub-Nyquist sampling method.

Introduction

Sub-Nyquist sampling as the basis for the digital coding of a composite PAL signal has been the subject of active investigation in the UK during the last two years by both the BBC and the IBA. The reasons for this interest arise from a rather complex interaction of conflicting requirements, long and short term, national and international, quality and cost. An interim culmination of this work was a demonstration of some experimental digital equipment presented to the EBU Technical Committee in Venice during April 1977, which, it was hoped, would lead to a submission to CCIR specifying common European standards for the digital coding of television signals. This article explains some of the technical background to the work, and attempts to clarify what, from a restricted viewpoint, might seem to be a surprising choice of sampling standard.

Background

The coding standard is required to meet the needs of the studio, and of both national and international links, while maintaining adequate quality. This simple statement contains the seeds of considerable debate. The solutions which have been suggested encompass data rates which extend from an uncompromising 213 Mbit/s to the 34 Mbit/s system required for international exchange. Sub-Nyquist PAL, requiring a data rate of 80 Mbit/s, falls in the lower half of the range. Some of the most important considerations of which account must be taken in making a choice are set out below.

The efficiency of the composite PAL signal in a one-wire analogue environment is unchallengeable. However, the system is inextricably tied to a series of minor impairments and inconveniences. These include the incidence of cross-colour, a slight loss of diagonal resolution and the problems of signal processing associated with the 8-field sequence. The PAL system was designed originally to overcome certain impairments of transmission which will not be significant in a digital environment. Moreover, the sampled-data approach lends itself naturally to a multiplexed component (YUV) format which would not suffer from

the disadvantages of PAL. The prospect of a fully digital studio therefore presents the opportunity of banishing PAL to its natural domain, that of analogue transmission, and using a component coding system. A number of European countries have additional domestic reasons for preferring the YUV form and have decided in its favour, but the UK broadcasters are resisting this conclusion. Why? The answer is that, on closer examination, the advantages of component coding in the UK prove to be largely illusory.

If the PAL system is to be retained for analogue transmission (and there is no other possibility in the foreseeable future) then the minor impairments of cross-colour and low-diagonal resolution will always be present for the domestic viewer. In this circumstance there is little value in eliminating PAL impairments in the studio because they will be re-introduced before transmission. Also, the advantages of signal processing are likely to be outweighed by short-term penalties which arise when introducing the component scheme.

In signal processing, the full advantages of component coding can be achieved only if the Y, U and V signals are sampled at a rate which is synchronised to both line frequency and frame frequency. Unfortunately, it is difficult to accommodate a frame-locked sampling pattern in a signal which is ultimately to be coded into PAL. The reason for this is that the sampling frequency of a PAL signal must be a multiple of sub-carrier, and the resulting pattern is asynchronous with frame frequency except at the high sample rate of $4f_{sc}$. Even in this case, only the luminance signal has samples which are picture-locked unless the data rate of the decoded YUV components is to exceed 213 Mbit/s. However, it is possible to accommodate a line-locked sample pattern by employing a slight change of picture geometry. This is discussed in more detail in the companion articles contained in this issue. The most desirable YUV formats are therefore largely incompatible with the most desirable PAL formats. The result is that there are significant difficulties in introducing an acceptable YUV standard within a PAL environment unless it be done by replacing complete studios, or by the introduction of a series of analogue PAL/YUV codecs. This latter course would be likely to produce an unacceptable level of impairment.

In addition to the arguments outlined above, the PAL signal format is considerably more efficient than that for YUV, demanding only two-thirds of the data rate. It therefore offers advantages in terms of line-transmission and tape-storage costs. In summary,

although UK broadcasters generally accept the long-term advantages of component coding, these are seen to be unattainable in the short term, while the immediate disadvantages of a transition to component coding are very apparent.

The early work on digital coding for composite PAL and NTSC was universally based on sampling at three times colour sub-carrier frequency ($3f_{sc}$). This had the positive advantage of being the lowest sub-carrier multiple that meets the Nyquist sampling criterion. In many ways this sampling rate seemed to be the optimum choice but, as will be shown, its use would cause a number of difficulties not easily overcome. The most important of these occurs in the processing of digital signals, particularly when it is required to convert between the YUV and composite forms as will be necessary in a number of studio situations and in the generation of signals for international exchange. Existing analogue coders and decoders introduce significant impairments, and near digital equivalents are not expected to be much superior although it may be possible to avoid the group-delay error caused by the sub-carrier notch. The reason for the PAL/YUV separation problem is the non-orthogonal sample pattern produced by $3f_{sc}$ sampling. This is illustrated in Fig. 1 on page 18.

Experience has shown that to achieve good results the conversion process should employ two-dimensional (comb) filtering techniques. With $3f_{sc}$ sampling, the methods available for filter design either result in infinite arrays of usually undesirable coefficients, or finite arrays for which the two-dimensional response is unsatisfactory. Even for one-dimensional filters the $3f_{sc}$ sample rate usually demands inconvenient coefficients. In principle, the $3f_{sc}$ sample pattern can be made to become orthogonal by shifting the phase of the clock pulses by 90° at the start of each line. In practice such methods are difficult to implement and are subject to phase error. The second difficulty with $3f_{sc}$ sampling is that the conversion to a twice (or near twice) sub-carrier sampling rate, which is presently envisaged for the luminance signal in France and Germany, requires an interface, the design and performance of which are problems of some magnitude. Finally, wherever it is desirable or necessary to reduce the bit rate below that attainable with linear $3f_{sc}$ pcm, the subjective performance of $3f_{sc}$ systems is inferior to the sub-Nyquist $2f_{sc}$ systems which are to be described.

A sub-Nyquist system based on sampling at $2f_{sc}$ was proposed by the BBC as a possible approach to bit-rate reduction. At that time it was not regarded

as a satisfactory studio standard because it was thought that the impairments introduced would be cumulative in cascaded systems. Subsequently, it was realised (and largely confirmed by simulation studies) that in principle this is not so and that the system was worthy of serious consideration. In particular, the system could deal with the conversion from PAL to a component form with relative ease. In this case samples of the component signals are locked to sub-carrier frequency.

A meeting of EBU Sub-Group C1 confirmed that the sampling frequencies proposed by member countries were very close, and that there was the possibility of an agreed standard. It was therefore suggested that a demonstration of possible standards should be given to the full EBU Technical Committee in the hope that, following this, a joint submission to CCIR proposing agreed standards for digital television and sound would be possible. The demonstration took place in Venice on 21st April 1977 and was supported by video contributions from IBA, ITCA, BBC, CCETT (France) and CBS (USA). The IBA and the BBC demonstrations were largely complementary. The BBC demonstrated a $2f_{sc}$ sub-Nyquist system using analogue comb filters, and a system which separated PAL into YUV components, while the IBA demonstrated a sub-Nyquist system using digital comb filters and, more significantly, an assembly of the basic elements of a digital television studio based on $2f_{sc}$ sampling. In addition, a multiplexed vision and sound transmission via the *Symphonie* satellite was provided by CCETT, digital noise reduction and special-effects mixers using frame stores were demonstrated on video tape by CBS, as was the technical performance of cascaded adc-dac codecs by ITCA.

The philosophy of the IBA demonstration was to provide not only an appraisal of the quality attainable with $2f_{sc}$ sub-Nyquist sampling, but to do so in a way such that its practical relevance within a studio complex would be apparent. The method by which this aim was to be achieved is shown schematically in Fig. 2 on page 28. In effect, the basic elements of a digital television studio were provided with the digital mixer as the interconnecting device. The individual equipments are considered in detail elsewhere in this edition of *IBA Technical Review*, and will not be further described here except to explain the reason for including the differential pcm module.

In both France and Germany it is envisaged that the composite signal will be converted to YUV form before digital coding. In France this choice is largely due to the characteristics of a composite SECAM signal which at present cannot be digitally coded. In Germany the component coding system was chosen because of the need to achieve a bit rate of 34 Mbit/s, which is the highest rate that their existing radio relay links can support. It was then thought that, for a given data rate, the highest quality would be achieved with component rather than composite coding. In the UK, where the situation encouraged the adoption of a composite standard, it was worthwhile investigating the potential for bit-rate reduction of sub-Nyquist PAL signals. A dpcm equipment was therefore included in the development programme and, in the event, achieved a picture quality which was as good as, if not better than, the more complex component coding system.

The $2f_{sc}$ composite signal lies at the heart of the standards proposed for the UK, and the remainder of this article deals with the theoretical foundations of sub-Nyquist PAL coding.
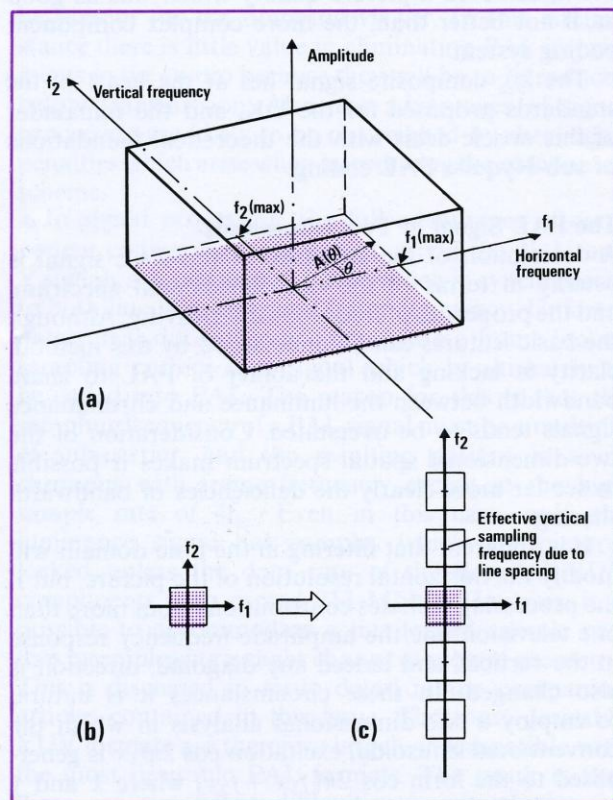
The PAL Signal in Two Dimensions

A description of the structure of the PAL signal is usually in terms of the one-dimensional spectrum, and the properties of frequency interleaving. Although the basic features can be understood by this method, clarity is lacking and the ability of PAL to share bandwidth between the luminance and chrominance signals tends to be overstated. Consideration of the two-dimensional spatial spectrum makes it possible to see far more clearly the deficiencies of bandwidth sharing.

It is apparent that filtering in the time domain will modify the horizontal resolution of the picture, but if the processing includes contributions from more than one television line the amplitude-frequency response in the vertical, and indeed any diagonal, direction is also changed. In these circumstances it is natural to employ a two-dimensional analysis in which the conventional sinusoidal excitation $\cos 2\pi f_1 x$ is generalised to the form $\cos 2\pi(f_1 x + f_2 y)$ where x and y respectively represent the horizontal and vertical axes of a television picture. A simple two-dimensional spectrum is shown in Fig. 1(a) to illustrate the method. Negative frequencies are included, as is the normal convention in the one-dimensional case. For the idealised filter shown, the characteristics can be described in terms of extinction frequencies, $f_1(max)$ and $f_2(max)$, for the horizontal and vertical directions respectively. For a spatial frequency $\cos 2\pi(f_1 x + f_2 y)$ at an angle θ to the horizontal, $\tan \theta = f_2/f_1$ and the extinction frequency is given by the magnitude of the

vector $A(\theta)$. Note that the maximum extinction frequency is $[f_1^2(\text{max}) + f_2^2(\text{max})]^{1/2}$ which, for equal horizontal and vertical bandwidths, is greatest for diagonal frequencies, and exceeds the horizontal bandwidth by a factor of $\sqrt{2}$. Idealised spectra of this kind can be drawn more easily in plan view, and this method will be used wherever possible.

Figure 1(b) shows a plan view of this two-dimensional spectrum, and Fig. 1(c) illustrates the result of the vertical sampling of the original scene by the process of line scanning. Replicas of the original spectrum appear centred on harmonics of the vertical sampling rate. If aliasing is to be avoided, these repeat spectra should not overlay the baseband signal.

A problem that occurs whenever it is necessary to consider the vertical resolution of television pictures is how best to encompass interlace. Fundamentally, the correct approach is to accept that, even for still pictures, an interlaced television raster is a three-dimensional signal, i.e., two spatial dimensions and one dimension of time. However, this greatly complicates the analysis and is not always necessary. The assumption upon which the success of interlace largely depends is that information conveyed by each field is retained unchanged, by the combined integration of the display tube phosphor and the eye, until the next field appears. If this integration were perfect, the system would have the full resolution of 312 cycles per picture height (c/ph) of which a 625-line system is capable. In practice, the integration is not perfect, and the interlaced scan has some of the properties of a system having only 312 lines, see Fig. 2. In particular, signals which contain vertical frequencies between 156 and 312 c/ph will cause a degree of aliasing. The effect is analysed in Appendix 1 where it is shown that for a subjective interfield decay of $e^{-\alpha}$ the relative magnitude of the repeat spectrum at 312 c/ph is $\tanh(\alpha/2)$. In this presentation the interlace problem will be circumvented by noting that spectral cancellation of the alias components is incomplete; this validates a treatment which considers each field separately.

The modulated sub-carrier is of course added to the one-dimensional signal, but it can be incorporated into the two-dimensional analysis by making use of
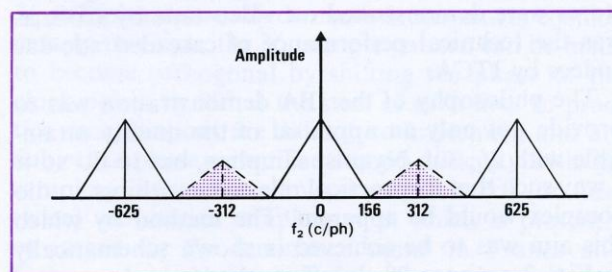

Fig. 1. For the analysis of signals which represent two-dimensional images, it is useful to employ a two-dimensional approach in the frequency domain. The figure at (*a*) shows a simple two-dimensional spectrum for which the bandwidths in the vertical and horizontal directions are equal. Often, a plan view of the frequency plane as at (*b*) is sufficient to bring out the essential features of two-dimensional spectra. The method is used here to show that the process of line-scanning effectively samples the scene in the vertical direction producing replicas of the baseband spectrum. This is shown at (*c*). The repeat spectra are centred on multiples of the vertical sampling frequency, i.e., the number of lines per unit distance on the display, and aliasing will occur at vertical frequencies greater than half this frequency.

Fig. 2. For an interlaced scan, an ambiguity arises concerning the vertical sampling frequency. If the field repetition rate is low, the eye will perceive each field separately, and the vertical sampling frequency will correspond to the number of lines in a single field. Alternatively, a high field repetition rate causes the eye to perceive multiple fields, and the number of lines is effectively doubled. The 20 ms field (containing $312\frac{1}{2}$ lines) used for television systems produces subjective effects belonging to both of these models. Vertical frequencies in the range 156–312 cpph are aliased in a single field, but the alias components are partially cancelled by the interlaced field which follows. The remaining (uncancelled) alias components are shown shaded.

the simple relationship that exists between spatial and temporal frequencies in a line-scanned system.[1] For a stationary scene, the temporal frequency is separated into integral and fractional multiples of line frequency. The integral multiple accounts for the horizontal component and the remainder accounts for the vertical component. For a single field of a PAL signal, the relationship is that a temporal frequency, f, gives rise to a spatial frequency of n cycles per picture width (c/pw), m cycles per picture height (c/ph):
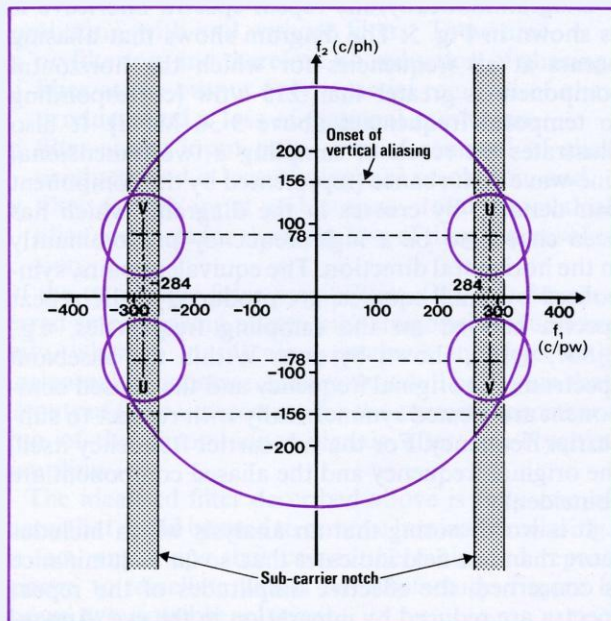
where
$$f = f_L(n - m/312 \cdot 5)$$

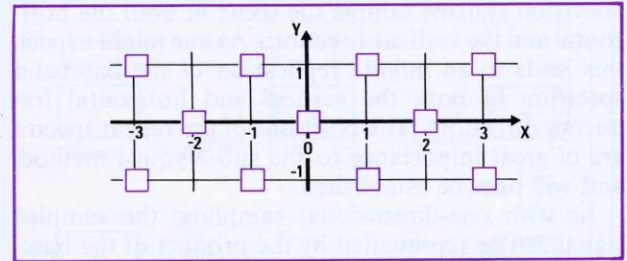$$f_L = \text{line frequency}$$

and
$$n, m = \text{integers}.$$

Application of the formula in the case of the U signal, which modulates sub-carrier frequency, gives

$$f_{sc} \simeq 15 \cdot 625 \, (284 - 78/312 \cdot 5) \text{ kHz}.$$

The U signal, therefore, occupies a region of the two-dimensional spectrum near 284 c/pw, 78 c/ph. The PAL switch introduced for the V component modi-

Fig. 4. In the sub-Nyquist system, the non-integral ratio between line frequency and sampling frequency causes the line-interleaved pattern of sampling sites shown above. The x and y axes have been scaled to simplify Fourier analysis.

fies the sub-carrier signal by a half-cycle per line, causing the V signal to separate from the U signal and to occupy a region of the spectrum near 284 c/pw, −78 c/ph. This is shown in Fig. 3 which also clearly demonstrates the limits of bandwidth sharing of the YUV components in PAL. Diagonal luminance frequencies in the region of the chrominance signals cause cross-colour and, with either one- or two-dimensional filtering, there is no way in which this can be prevented without impairing luminance resolution. It follows that these diagonal luminance frequencies are of little value, and their removal, or attenuation, will probably enhance rather than degrade subjective performance. The conventional PAL coder is often provided with a notch at sub-carrier frequency in recognition of this fact, but the notch, being a one-dimensional filter, reduces the amplitude of not only the diagonal frequencies but all frequencies in the shaded regions of Fig. 3. Of particular importance is the loss of horizontal and near horizontal frequencies since there is some evidence to suggest that in typical pictures there is a preponderance of horizontal and vertical frequencies over diagonal frequencies.

In summary it can be said that the PAL system cannot adequately reproduce certain diagonal luminance frequencies, and that systems which degrade the resolution in these areas of the spectrum are to be preferred to those which degrade resolution in either the horizontal or vertical directions. This is an important conclusion to be taken into account in an assessment of the sub-Nyquist sampling technique.

Fig. 3. In the PAL system, the U and V chrominance signals occupy regions of the two-dimensional spectrum, shown by the small circles, which are required for the reproduction of certain diagonal luminance frequencies. Because these frequencies will cause cross-colour effects they are sometimes excluded in the PAL coding process by using a notch filter in the luminance channel. The notch filter attenuates all luminance frequencies in the shaded areas of the diagram, and therefore removes horizontal frequencies which could be retained in a more sophisticated design of coder.

Sub-Nyquist Luminance

It has been seen that sampling a scene in the vertical direction with a line scanning process leads to an infinite replication of the two-dimensional baseband spectrum along the vertical frequency axis. All digital

television systems sample the scene in both the horizontal and the vertical directions. As one might expect, this leads to an infinite replication of the baseband spectrum in both the vertical and horizontal frequency directions. The positions of the repeat spectra are of great importance to the sub-Nyquist method, and will now be established.

As with one-dimensional sampling, the sampled signal can be represented by the product of the baseband signal and the sampling pattern:

$$F_s(x, y) = F(x, y) . S(x, y) \quad . . \quad . . \quad (1)$$

The sampling pattern consists of a series of impulse functions located at the sampling sites. In the case of $2f_{sc}$ sampling there are almost exactly $567\frac{1}{2}$ samples per television line, and the non-integral ratio causes an interleaved pattern as shown in Fig. 4. A scaling

factor has been introduced in this figure to simplify the analysis. The regular sampling pattern may be represented by a two-dimensional Fourier series as follows:

$$S(x, y) = \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} C_{p,q} e^{j\pi(px + qy)} \quad . . \quad (2)$$

In this form it is clear that the sampling process of Equation (1) causes the baseband spectrum to modulate each of the sinusoids $e^{j\pi(px + qy)}$ resulting in a repeat spectrum of magnitude $C_{p,q}$ on the frequency site p, q. (In the normalised form, the site 1, 1 corresponds to the spatial frequency 567 c/pw, 156 c/ph.) It follows that the two-dimensional spectrum of the sampled signal can be established merely by calculating the coefficient $C_{p,q}$ using a Fourier integration formula. This method has been used in Appendix 2 for the case of the sub-Nyquist sampling pattern of Fig. 4. The result is:

$$C_{p,q} = 1, \text{ for } p + q \text{ even,}$$
$$= 0, \text{ for } p + q \text{ odd.}$$

Diagrammatically, the repeat spectra interleave is as shown in Fig. 5. The diagram shows that aliasing occurs at all frequencies for which the horizontal component is greater than 215 c/pw (corresponding to temporal frequencies above 3·36 MHz). It also illustrates the result of sampling a two-dimensional sine-wave disturbance (represented by the component pair denoted by crosses in the diagram) which has been chosen to be a high frequency predominantly in the horizontal direction. The equivalent pairs, symbolised by small squares, are produced in the repeat spectra centred on the sampling frequencies, e.g., ± 567, ± 156, shown by dots. Within the baseband spectrum, the original frequency and the aliased component are located symmetrically with respect to sub-carrier frequency. For the sub-carrier frequency itself, the original frequency and the aliased component are coincident.

It is worth noting that an analysis which includes more than one field indicates that, so far as luminance is concerned, the effective amplitudes of the repeat spectra are reduced by integration in the eye. Appendix 3 shows that, if the subjective decay between successive fields is $e^{-\alpha}$, then the apparent amplitude of repeat luminance spectra is $(1 - \text{sech}\,\alpha)^{1/2}$. The cancellation is, however, incomplete and the alias components must be attenuated in the reconstruction of the luminance signal. The two-dimensional filter shown in Fig. 6 would be effective in the complete removal of the unwanted components, and Fig. 7
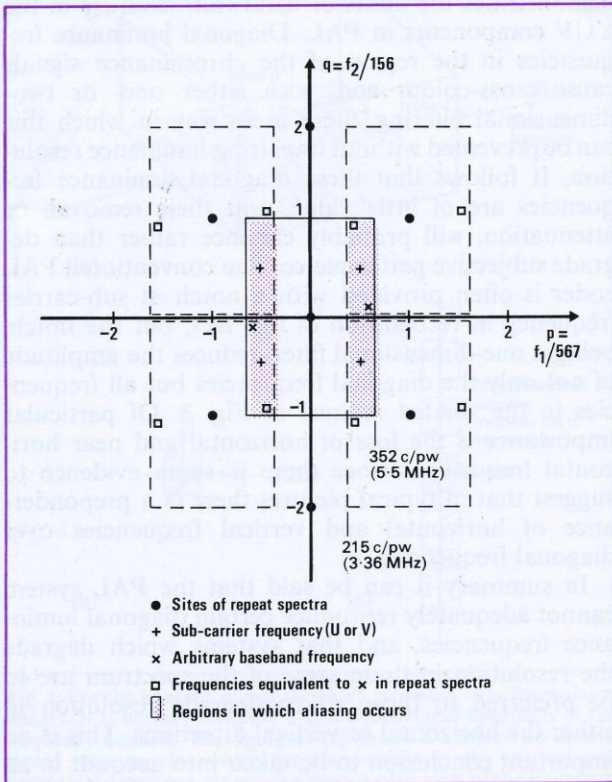
Fig. 5. In digital television systems, the scene is sampled in both the horizontal and vertical directions which causes a replication of the baseband spectrum at points throughout the frequency plane. In general, the sampling rates are so chosen that, for frequencies within the bandwidth specifications, there is no aliasing between the baseband and the repeat spectra. However, for $2f_{sc}$ sub-Nyquist sampling, these spectra overlap in the manner shown, and it is necessary to modify the shape of the baseband spectrum prior to the sampling process, see Fig. 6.

Fig. 6. If the idealised filter shown here were used both before and after the $2f_{sc}$ sampling process it would totally eliminate luminance aliasing.

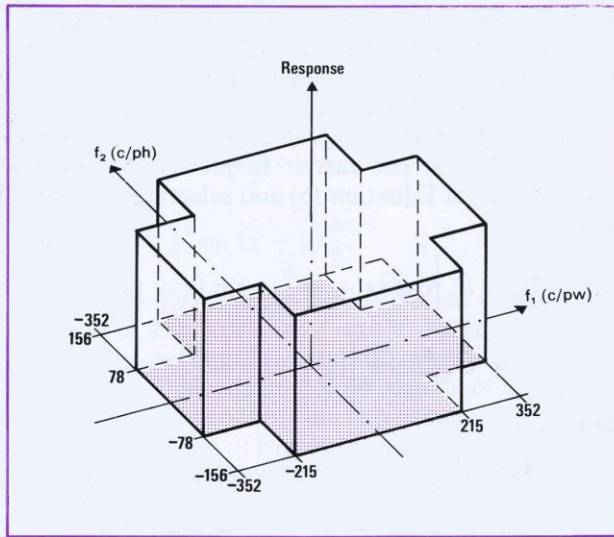**Fig. 7.** The figure illustrates the possible outcomes of using the filter of Fig. 6 before sampling and/or in the reconstruction of the baseband spectrum (see text). The use of a single filter at either input or output results in a degree of luminance aliasing; this can be eliminated by including filters in both positions.

illustrates the effects of operating in the four possible conditions, with and without filters. These are:

i no filtering, and therefore no reduction of aliasing,

ii filter at the output only—partial removal of alias products and a loss of diagonal resolution,

iii filter at the input only—partial removal of alias products and a loss of diagonal resolution, and

iv filter at both input and output—total removal of alias products for the same loss of diagonal resolution.

Of the two single-filter possibilities, (iii) is to be preferred since aliasing affects high diagonal frequencies only, whereas in (ii) alias products interfere with horizontal frequencies. However, it will be shown that an output filter is necessary for the correct reconstruction of the chrominance signals after sub-Nyquist sampling.

The idealised filter described above is only one of many that could be used to remove luminance aliasing. In fact, any shape that causes the main and repeat spectra to tessellate would be satisfactory. Figure 8 shows two possible solutions.

Sub-Nyquist Chrominance

For the chrominance signals there can be no question of subjective cancellation between fields because they are displayed as new baseband spectra after decoding. The decoding process uses information from one field only, and the alias components therefore have their full effect.

Fig. 8. The filter characteristic of Fig. 6 is not the only one capable of eliminating luminance alias products. Numerous characteristics are possible, all of which satisfy certain symmetry constraints which cause the spectra to tessellate in the frequency domain. Two filters are shown which would satisfy the requirements of the luminance spectrum, but additional constraints must be introduced to ensure the proper reconstruction of the chrominance signals.

Using the normalised coordinates of Fig. 4, the chrominance signals in an area of uniform colour may be represented by the spatial frequency terms:

$$U = \hat{U} \cos (x + y) \frac{\pi}{2} \quad \dots \dots \quad (3)$$

and

$$V = \hat{V} \sin (x - y) \frac{\pi}{2} \quad \dots \dots \quad (4)$$

These functions are of course in phase quadrature along the x axis. Appendix 4 shows the effect of sampling the chrominance signals with a sub-Nyquist pattern of arbitrary phase. The result is the addition of an alias term to each of the chrominance components as follows:

$$U_s = \hat{U} \left[\cos (x + y) \frac{\pi}{2} + \cos \left\{ (x + y) \frac{\pi}{2} + \phi \right\} \right], \quad (5)$$

and

$$V_s = \hat{V} \left[\sin (x - y) \frac{\pi}{2} - \sin \left\{ (x - y) \frac{\pi}{2} \pm \phi \right\} \right], \quad (6)$$

The alias components will distort the chrominance signals in both amplitude and phase. However, by

Fig. 9. Sampling the chrominance signals at the $2f_{sc}$ rate causes a 3 dB increase in the level of the sub-carrier. In addition, the quadrature relationship between the U and V components is lost in the sampling process. Reconstruction of the chrominance information therefore requires a 3 dB attenuation of the chrominance signals, and the introduction of a vertical phase shift which restores the correct U and V sub-carrier relationship. The figure shows the ideal amplitude response for the chrominance reconstruction filter. For convenience only two of the four points representing sub-carrier frequency are indicated.

careful choice of the sampling phase, ϕ, the amplitude ratio of the U and V signals may be preserved. Two values of sampling phase are capable of this, namely $\phi = \mp \pi/2$. The phase angle $\pi/2$ is referred to the sampling frequency $2f_{sc}$, and therefore corresponds to a $\pi/4$ shift at sub-carrier frequency. Taking the negative sign in Equation (6) and selecting $\phi = +\pi/2$ gives:

$$U_s = \hat{U} \left[\cos (x + y) \frac{\pi}{2} - \sin (x + y) \frac{\pi}{2} \right]$$

$$= \hat{U} \sqrt{2} \cos \left[(x + y) \frac{\pi}{2} + \frac{\pi}{4} \right], \quad \dots \dots \quad (7)$$
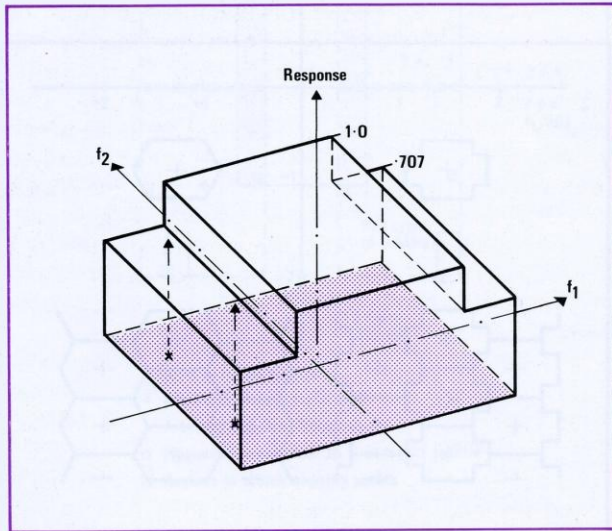
and

$$V_s = \hat{V} \left[\sin (x - y) \frac{\pi}{2} - \cos (x - y) \frac{\pi}{2} \right]$$

$$= -\hat{V} \sqrt{2} \cos \left[(x - y) \frac{\pi}{2} + \frac{\pi}{4} \right]. \quad \dots \quad (8)$$

The ratio between the U and V amplitudes has been preserved, but both have been increased in level by a factor of $\sqrt{2}$. Moreover, the quadrature relationship [Equations (3) and (4)] has been lost and the U and V sub-carrier signals are now in antiphase along the x axis. This is a fundamental characteristic of $2f_{sc}$ sampling; to reconstruct the chrominance signals it is necessary to introduce an attenuation of 3 dB, and to restore the quadrature relationship between the U and V sub-carrier signals. It may be shown that the reconstruction requirements are the same for areas of non-uniform colour, which cause sidebands symmetrically displaced from the sub-carrier frequency. The ideal amplitude characteristic for the chrominance reconstruction filter is therefore as shown in Fig. 9.

The idealised luminance and chrominance output filters of Figs. 6 and 9 are mutually incompatible, and some compromise must be sought. The restoration of the correct chrominance phase relationship is central to the solution adopted. Returning to Equations (7) and (8), if two adjacent lines of sub-Nyquist chrominance, corresponding to y and $y - 1$, are averaged, the result is:

for the U signal,

$$U = \sqrt{2} \frac{\hat{U}}{2} \left\{ \cos \left[(x + y) \frac{\pi}{2} + \frac{\pi}{4} \right] \right.$$

$$\left. + \cos \left[(x + y - 1) \frac{\pi}{2} + \frac{\pi}{4} \right] \right\}$$

$$= \hat{U} \cos (x + y) \frac{\pi}{2},$$

and, for the V signal,
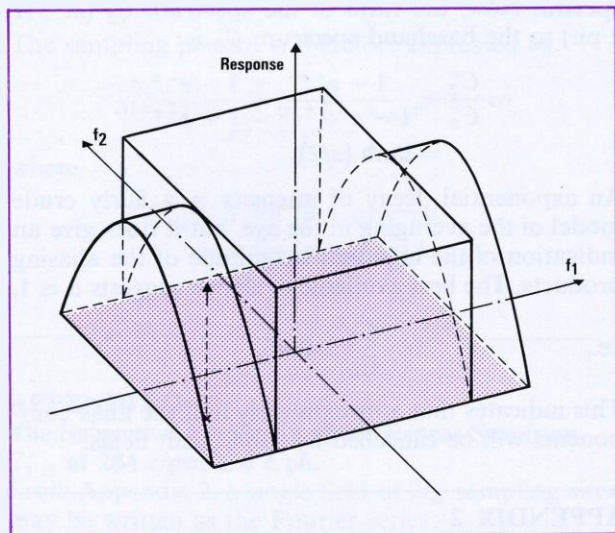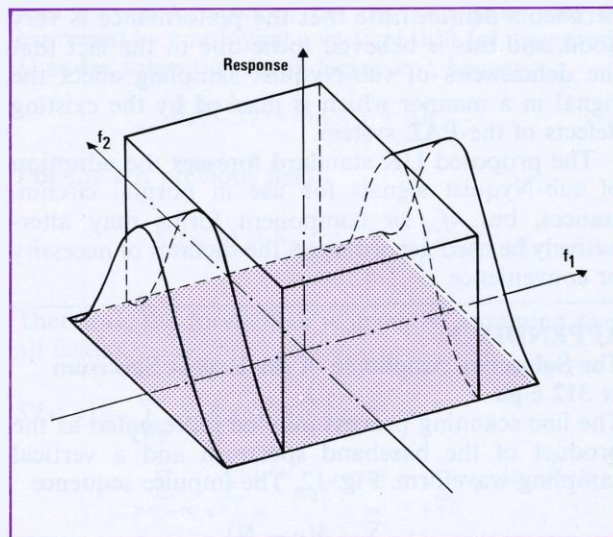
$$V = -\sqrt{2}\,\frac{\hat{V}}{2}\left\{\cos\left[(x-y)\frac{\pi}{2}+\frac{\pi}{4}\right]\right.$$
$$\left.+\cos\left[(x-y+1)\frac{\pi}{2}+\frac{\pi}{4}\right]\right\}$$
$$= \hat{V}\sin(x-y)\frac{\pi}{2}.$$

Spatial frequencies having horizontal components in excess of 215 c/pw, i.e., $f = 3.36$ MHz, are reconstructed by averaging two sub-Nyquist lines, thereby regenerating the original colour signals of Equations (3) and (4). This produces the two-dimensional amplitude characteristic of Fig. 10, and results in a reasonable compromise between the ideal luminance and chrominance filters (Figs. 6 and 9).

The description given demonstrates the reconstruction of a chrominance signal which has been sampled at $2f_{sc}$. It has already been established that a pre-sampling filter is necessary (for the attenuation of alias products), and it might be thought that this filter would modify the chrominance signals in a manner which would invalidate the analysis of chrom-

Fig. 11. The diagram shows the overall luminance response when filters of the type shown in Fig. 10 are used at the input and output of a sub-Nyquist system. Although the characteristics are less than ideal for both luminance and chrominance, the impairments which result are very difficult to observe in practice.

inance reconstruction. In fact, that is not the case. It can be shown that, if a pre-sampling filter is used which is identical to the reconstruction filter described, it will have no effect on the sub-Nyquist chrominance signals. The reason for this apparent paradox is that the sampling process generates alias components which entirely replace the chrominance losses. This produces sub-carrier signals identical to those which would have been present if the pre-sampling filter had not been introduced. The manner in which this occurs is explained in Appendix 5. The pre-sampling filter is, however, effective in reducing luminance aliasing, and when used in combination with an identical reconstruction filter the overall luminance response is as shown in Fig. 11.

Filters of the type described cannot completely remove alias products, nor can they faithfully reconstruct the chrominance sidebands. However, in practice it is almost impossible to discern these imperfections. The only detectable impairment is a slight loss of diagonal luminance resolution.

Conclusion

An attempt has been made to show the advantages and disadvantages of sub-Nyquist sampling of a composite PAL signal. The performance is best described by a two-dimensional presentation which illustrates more clearly the relationship between the signal processing and the subjective effects which result. Ex-

Fig. 10. The ideal filter characteristics for the restoration of the luminance and chrominance signals are shown respectively in Figs. 6 and 9. These characteristics are incompatible, and the filter characteristic used in practice represents a compromise between the two. The characteristic shown here is formed by averaging adjacent sub-Nyquist lines for temporal frequencies above 3.36 MHz. This restores the correct sub-carrier relationships as well as forming a reasonable compromise between the conflicting requirements for the two amplitude responses. For simplicity only one of the four points representing sub-carrier frequency is shown.

11

periments demonstrate that the performance is very good, and this is believed to be due to the fact that the deficiencies of sub-Nyquist sampling affect the signal in a manner which is masked by the existing defects of the PAL system.

The proposed UK standard foresees the adoption of sub-Nyquist signals for use in normal circumstances, but $4f_{sc}$ or component forms may alternatively be used according to the dictates of necessity or convenience.

APPENDIX 1
The Subjective Amplitude of the Repeat Spectrum at 312 c/ph

The line scanning process may be represented as the product of the baseband spectrum and a vertical sampling waveform, Fig. 12. The impulse sequence

$$\sum_{N=-\infty}^{\infty} \delta(y - N)$$

may be expanded as a Fourier series, thus,

$$F_0(y) = \sum_{k=-\infty}^{\infty} C_k e^{jk\pi y},$$

where
$$C_k = 1, \text{ for } k \text{ even},$$
$$= 0, \text{ for } k \text{ odd}.$$

The coefficients C_k determine the amplitude of the repeat spectra. The co-ordinates have been normalised so that the first repeat spectrum at $k = 2$ corresponds to a vertical frequency of 312 c/ph.

The scanning waveform in the second field is found by applying a vertical shift (to $y = \frac{1}{2}$) and a subjective attenuation $e^{-\alpha}$:

$$F_1(y) = F_0(y - \tfrac{1}{2})e^{-\alpha}$$
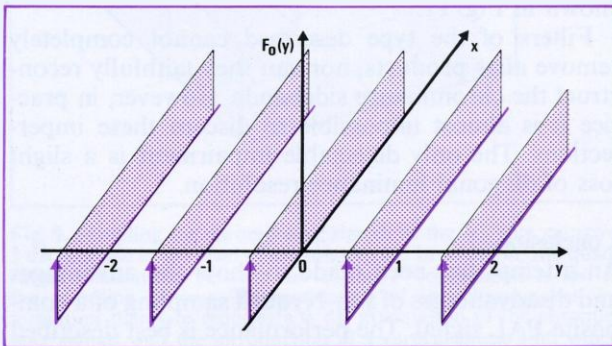$$= \sum_{k=-\infty}^{\infty} C_k e^{jk\pi(y - 1/2)}e^{-\alpha}.$$

Fig. 12. Line scanning can be considered as a vertical sampling process in which the sampling waveform is a series of impulses in the y direction extending indefinitely in the x direction.

The effect of the mth field scan is:

$$F_m(y) = \sum_{k=-\infty}^{\infty} C_k e^{jk\pi(y - m/2)}e^{-m\alpha}$$
$$= \sum_{k=-\infty}^{\infty} C_k e^{jk\pi y}e^{-m(\alpha + jk\pi/2)}.$$

The total effect is given by summing over all fields:

$$F'(y) = \sum_{k=-\infty}^{\infty} \sum_{m=0}^{\infty} C_k e^{jky\pi}e^{-m(\alpha + jk\pi/2)}$$
$$= \sum_{k=-\infty}^{\infty} C_k e^{jk\pi y} \sum_{m=0}^{\infty} e^{-m(\alpha + jk\pi/2)}.$$

The sum of the second series is $[1 - e^{-(\alpha + jk\pi/2)}]^{-1}$,

$$\therefore \quad F'(y) = \sum_{k=-\infty}^{\infty} \frac{C_k}{[1 - e^{-(\alpha + jk\pi/2)}]} \cdot e^{jk\pi y}$$
$$= \sum_{k=-\infty}^{\infty} C'_k e^{jk\pi y},$$

where
$$C'_k = \frac{C_k}{[1 - e^{-(\alpha + jk\pi/2)}]}.$$

The coefficients C'_k are the amplitudes of the repeat spectra. Now, the ratio of the spectrum C'_2 (at 312 c/ph) to the baseband spectrum C'_0 is:

$$\frac{C'_2}{C'_0} = \frac{1 - e^{-\alpha}}{1 - e^{-(\alpha + j\pi)}} = \frac{1 - e^{-\alpha}}{1 + e^{-\alpha}}$$
$$= \tanh(\alpha/2).$$

An exponential decay of intensity is a fairly crude model of the averaging in the eye, but it does give an indication of the apparent magnitude of the aliasing products. The best available evidence suggests $\alpha \simeq 1$,

i.e.,
$$\frac{C'_2}{C'_0} \simeq 0.46.$$

This indicates that approximately half the alias components will be cancelled by subsequent fields.

APPENDIX 2
Spatial Frequencies Present in a $2f_{sc}$ Sampling Pattern

The sampling pattern is represented by an array of impulse functions on the sites shown in Fig. 4. The pattern may be expressed as the two-dimensional Fourier series:

$$S(x, y) = \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} C_{p,q} e^{jp\pi x}e^{jq\pi y},$$

where

$$C_{p,q} = \frac{1}{4} \int_{-1}^{1} \int_{-1}^{1} S(x, y) e^{-jp\pi x} e^{-jq\pi y} \, dx \, dy$$

and

$$S(x, y) = 2\delta(x - n, y - m), \text{ for } n + m \text{ even,}$$
$$= 0, \text{ for } n + m \text{ odd.}$$

The factor of 2 is introduced to maintain a unity coefficient for the baseband spectrum. Therefore, for $n + m$ even:

$$C_{p,q} = \frac{1}{2} \int_{-1}^{1} \int_{-1}^{1} \delta(x - n, y - m) e^{-j\pi(px + qy)} \, dx \, dy$$
$$= \frac{1}{2} [1 + \frac{1}{4} \{ e^{-j\pi(p+q)} + e^{j\pi(p+q)}$$
$$+ e^{-j\pi(p-q)} + e^{j\pi(p-q)} \}].$$

This may be rearranged to give:

$$C_{p,q} = \frac{1}{2} [1 + \cos p\pi \cos q\pi]$$
$$= \frac{1}{2} [1 + (-1)^{p+q}],$$

i.e.,

$$C_{p,q} = 1, \text{ for } p + q \text{ even,}$$
$$= 0, \text{ for } p + q \text{ odd.}$$

The sampling pattern is therefore expressed as:

$$S(x, y) = \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} C_{p,q} e^{j\pi(px + qy)},$$

where

$$C_{p,q} = 1, \text{ for } (p + q) \text{ even,}$$
$$= 0, \text{ for } (p + q) \text{ odd.}$$

APPENDIX 3
The Subjective Amplitude of the Repeat Spectrum $C'_{1,1}$ at 284 c/pw, 156 c/ph.
From Appendix 2, a single field of $2f_{sc}$ sampling sites may be written as the Fourier series:

$$S_0 = \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} C_{p,q} e^{j\pi(px + qy)},$$

where

$$C_{p,q} = 1, \text{ for } p + q \text{ even,}$$
$$= 0, \text{ for } p + q \text{ odd.}$$

The subjective effect of the previous field may be expressed by employing a vertical shift (of magnitude $\frac{1}{2}$) and a subjective decay factor $e^{-\alpha}$, hence:

$$S_1(x, y) = S_0(x, y + \frac{1}{2}) e^{-\alpha}.$$

For the kth previous field,

$$S_k(x, y) = S_0 \left(x, y + \frac{k}{2} \right) e^{-k\alpha}.$$

Therefore, the total effect is given by summing over all fields:

$$S'(x, y) = \sum_{k=0}^{\infty} S_0 \left(x, y + \frac{k}{2} \right) e^{-k\alpha}$$
$$= \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} C_{p,q} e^{j\pi(px + qy)} \sum_{k=0}^{\infty} e^{-k(\alpha - jq\pi/2)}.$$

The final series is summable giving:

$$S'(x, y) = \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} \frac{C_{p,q}}{1 - e^{-(\alpha - jq\pi/2)}} \cdot e^{j\pi(px + qy)}$$
$$= \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} C'_{p,q} e^{j\pi(px + qy)},$$

where

$$C'_{p,q} = C_{p,q} [1 - e^{-(\alpha - jq\pi/2)}]^{-1}.$$

Now, the ratio of the repeat spectrum $C'_{1,1}$ to the baseband spectrum $C'_{0,0}$ is:

$$\frac{|C'_{1,1}|}{|C'_{0,0}|} = \frac{1 - e^{-\alpha}}{|1 - e^{-(\alpha - j\pi/2)}|} = \frac{1 - e^{-\alpha}}{\sqrt{1 + e^{-2\alpha}}},$$
$$\therefore \quad \frac{|C'_{1,1}|}{|C'_{0,0}|} = (1 - \operatorname{sech} \alpha)^{1/2}.$$

If α is assumed to be in the region of unity, the relative magnitude of the repeat spectrum $C'_{1,1}$ is approximately 0·6.

APPENDIX 4
Sub-Nyquist Sampling of the Chrominance Signals
In an area of constant colour, the sub-carrier signal may be represented by the sum of two spatial frequencies:

$$U = \hat{U} \cos (x + y) \frac{\pi}{2},$$

$$V = \hat{V} \sin (x - y) \frac{\pi}{2}.$$

These signals are sampled by the two-dimensional pattern:

$$S(x, y) = \sum_{p=-\infty}^{\infty} \sum_{q=-\infty}^{\infty} C_{p,q} e^{j\pi(px+qy)}$$

$$= 1 +$$

$$2 \sum_{k=1}^{\infty} \sum_{l=1}^{\infty} [\cos (kx + ly)\pi + \cos (kx - ly)\pi].$$

Only terms for which $k = l = 1$ cause aliasing products which fall within the baseband spectrum.

For the U signal, the important terms are

$$1 + 2 \cos (x + y)\pi,$$

and for the V signal

$$1 + 2 \cos (x - y)\pi.$$

Let the sampling pattern have an arbitrary phase shift ϕ in the x or the y direction. Then, within the baseband spectrum:

$$U_s = \hat{U} \cos (x + y) \frac{\pi}{2}$$
$$\times \{1 + 2 \cos [(x + y)\pi + \phi]\}, \quad (A4.1)$$

$$V_s = \hat{V} \sin (x - y) \frac{\pi}{2}$$
$$\times \{1 + 2 \cos [(x - y)\pi \pm \phi]\}. \quad (A4.2)$$

Note the arbitrary sign for phase in Equation (A4.2). The sign depends on whether the phase shift is in the x or the y direction. Rearranging Equation (A4.1) gives:

$$U_s = \hat{U} \left\{ \cos (x + y) \frac{\pi}{2} + \cos \left[(x + y) \frac{\pi}{2} + \phi \right] \right.$$
$$\left. + \cos \left[(x + y) \frac{3\pi}{2} + \phi \right] \right\}.$$

The final term may be dropped since this falls outside the baseband spectrum, hence:

$$U_s = \hat{U} \left\{ \cos (x + y) \frac{\pi}{2} + \cos \left[(x + y) \frac{\pi}{2} + \phi \right] \right\}. \quad (A4.3)$$

A similar treatment of the V signal yields:

$$V_s = \hat{V} \left\{ \sin (x - y) \frac{\pi}{2} - \sin \left[(x - y) \frac{\pi}{2} \pm \phi \right] \right\}. \quad (A4.4)$$

In each case, there occurs an aliased term of the same spatial frequency, and the same amplitude, which will cause distortion of both the amplitude and phase of the chrominance signals. If the ratio of the U_s and V_s amplitudes are to remain the same as in the original signal, then the sampling phase, ϕ, must be chosen with care. (If $\phi = 0$, the V_s signal disappears altogether.) Examination of Equations (A4.3) and (A4.4) shows that there are only two possible values of ϕ which preserve the amplitude ratio U_s/V_s. These are $\phi = \mp (\pi/2)$, where the sign is the reverse of that chosen in Equation (A4.4). (A $\pi/2$ shift at the sampling frequency corresponds to a 45° shift at sub-carrier frequency.) If the negative sign is taken, then $\phi = \pi/2$ and Equations (A4.3) and (A4.4) simplify to:

$$U_s = \hat{U} \left[\cos (x + y) \frac{\pi}{2} - \sin (x + y) \frac{\pi}{2} \right]$$
$$= \hat{U} \sqrt{2} \cos \left[(x + y) \frac{\pi}{2} + \frac{\pi}{4} \right],$$

$$V_s = \hat{V} \left[\sin (x - y) \frac{\pi}{2} - \cos (x - y) \frac{\pi}{2} \right]$$
$$= -\hat{V} \sqrt{2} \cos \left[(x - y) \frac{\pi}{2} + \frac{\pi}{4} \right].$$

Note that the amplitudes of the two chrominance signals have increased by a factor of $\sqrt{2}$, also that the quadrature relationship between the U and V signals has been destroyed. The U_s and V_s signals are now in anti-phase along the x axis.

APPENDIX 5
The Effect of the Pre-Sampling Filter

It is required to show that, when the chrominance reconstruction filter is used as a pre-sampling filter, the sub-Nyquist chrominance signals are identical with those of Equations (7) and (8), i.e.:

$$U_s = \hat{U} \sqrt{2} \cos \left[(x + y) \frac{\pi}{2} + \frac{\pi}{4} \right],$$

and
$$V_s = -\hat{V} \sqrt{2} \cos \left[(x - y) \frac{\pi}{2} + \frac{\pi}{4} \right].$$

The chrominance input signals are described by the equations:

$$U = \hat{U} \cos (x + y) \frac{\pi}{2},$$

and
$$V = \hat{V} \sin (x - y) \frac{\pi}{2}.$$

The pre-sampling filter first averages two PAL video lines corresponding to y and $y + 1$, as follows:

$$U' = \frac{\hat{U}}{2}\left[\cos(x + y)\frac{\pi}{2} + \cos(x + y + 1)\frac{\pi}{2}\right]$$

$$= \frac{\hat{U}}{\sqrt{2}}\cos\left[(x + y)\frac{\pi}{2} + \frac{\pi}{4}\right],$$

and

$$V' = \frac{\hat{V}}{2}\left[\sin(x - y)\frac{\pi}{2} + \sin(x - y - 1)\frac{\pi}{2}\right]$$

$$= -\frac{\hat{V}}{\sqrt{2}}\cos\left[(x - y)\frac{\pi}{2} + \frac{\pi}{4}\right].$$

These signals are now sampled at $2f_{sc}$, corresponding to the $\pi/4$ and $5\pi/4$ phases of sub-carrier. The important spatial frequency terms in the sampling pattern are:

for the U' signal, $1 + 2\cos\left[(x + y)\pi + \frac{\pi}{2}\right]$,

and

for the V' signal, $1 + 2\cos\left[(x + y)\pi + \frac{\pi}{2}\right]$.

The baseband and alias terms are therefore:

$$U'_s = \frac{\hat{U}}{\sqrt{2}}\cos\left[(x + y)\frac{\pi}{2} + \frac{\pi}{4}\right]$$

$$\times \left[1 + 2\cos\left\{(x + y)\pi + \frac{\pi}{2}\right\}\right]$$

$$= \sqrt{2}\,\hat{U}\cos\left[(x + y)\frac{\pi}{2} + \frac{\pi}{4}\right],$$

and

$$V'_s = -\frac{\hat{V}}{\sqrt{2}}\cos\left[(x - y)\frac{\pi}{2} + \frac{\pi}{4}\right]$$

$$\times \left[1 + 2\cos\left\{(x - y)\pi + \frac{\pi}{2}\right\}\right]$$

$$= -\sqrt{2}\,\hat{V}\cos\left[(x - y)\frac{\pi}{2} + \frac{\pi}{4}\right].$$

These are identical with those of Equations (7) and (8), which are the unfiltered sub-Nyquist sub-carrier signals.

Reference
1. Hertz, P, and Gray, F, 'A Theory of Scanning and its Relation to the Transmitted Signal in Tele-photography and Television', *The Bell System Technical Journal*, 1934, 13, p. 464 et seq.

Proposed Digital Television Standards for 625-Line PAL Signals

by J L E Baldwin and I R Lever

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Synopsis

Digital techniques are moving rapidly into virtually all sectors of picture signal processing and transmission. The era of 'stand alone' digital devices employed within an analogue environment is nearing its end. As digital methods find their way into more areas of processing, the desirability of interconnecting between these areas, without converting between the digital and analogue forms, is obvious. The interconnections must also be digital.

For this to be possible the various parameters of the digital signal must be defined, specified and agreed not only nationally but also, hopefully, internationally, and by as early a date as possible.

Introduction

In real terms, the cost of digital processing has progressively decreased during the past ten years. If this trend continues, in a few more years it will result in digital television processing becoming generally cheaper than could be realised by analogue means. Already digital processing has become the norm in standards converters, synchronisers and video tape time-base correctors; in these applications it has par-ticular advantages and is also fully cost effective. It is now being introduced into telecines and special-effects mixers. Soon video tape recorders will become digital, and, perhaps last of all, the camera will follow suit.

At present these items are islands of digital processing in an analogue sea. The passage of a signal through each island requires conversion from analogue to digital form and back again. Such a pair of conversions, usually referred to as a codec, causes

16

a small impairment as a result of quantisation, and, where 8-bit words are used in the digital equipment, it is generally agreed that four in cascade cause a just perceptible impairment on critical material. Since the production of programmes often requires the passage of a signal through several video tape recorders, it follows that if these use digital processing, e.g., time-base correction, the amount of impairment introduced from this cause alone is becoming significant.

As time passes, more digital islands will progressively appear until they form an archipelago. Each island will add its contribution of quantising noise, each will need black-level clamps and some means of deriving clock pulses; each will add its own contribution to impairment.

This insidious degradation of picture quality must be avoided by constructing bridges from one digital island to the next so that the signal may pass directly from one digital equipment to another without the need for converting from and to analogue. This can be achieved only if the digital standards adopted in different items of equipment are compatible. Either a single standard, or at least a set of compatible standards, must be established at an early date in order that the equipment might not have to be scrapped within its normal life time.

Component versus Composite Coding

The coding of the component luminance and colour-difference signals has a natural attraction because, since it is independent of the choice of colour standard, problems of transcoding between PAL and SECAM would disappear. Is this fundamental attractiveness realisable in practice? Could equipment using component coding be introduced into existing PAL studio centres? This last question includes not only the studio centres in countries using PAL, but also those in which the origination occurs in PAL followed by a transcoding to SECAM prior to broadcasting. The standard chosen should not prevent the practical realisation of digital equivalents to analogue equipment, and it should facilitate, insofar as can be foreseen, the realisation of new types of signal processing. It must also be capable of interfacing with conventional analogue equipment without causing significant impairment.

It is well known that a pair of conventional PAL analogue coders and decoders do cause significant impairment. This is largely due to the one-dimensional filtering used to allegedly separate luminance and chrominance components in the decoder. It may be greatly improved by using two-dimensional filtering

for the separation. The stability of the 1-line delays that are required is very exacting both in terms of delay and gain, and it seems unlikely that these requirements could be maintained unless digital processing be used. Digital processing would be impracticable for this application unless sampling were at a frequency simply related to sub-carrier. In this case, of course, the digital signal leaving the analogue-to-digital converter is in the composite PAL form, and it becomes component coded only after two-dimensional filtering and the demodulation of the chrominance information. The component-coded form would thus also have word frequencies simply related to sub-carrier. For normal application the luminance word frequency would probably be twice that of sub-carrier while for chrominance it would be equal to sub-carrier. For other purposes it would be essential to use the composite-coded digital form, e.g., the application to modulators described by F H Wise.[1]

The conclusion that the sampling frequencies for component coding must be simply related to sub-carrier[2] is important. It is precisely the same relationship as would be desirable for composite coding.* The advantages usually claimed for the use of component coding are nearly all connected with the use of frame stores, and are that the processes of noise reduction and special effects, e.g., zoom and slow motion, are made easier. An exception is that video-tape editing is easier with component coding, whether or not a frame store is used.

Noise reduction is normally most important for electronic news-gathering (eng) types of operation. In this case the incoming picture is likely to have a composite form which implies that any saving made by using component coding for the noise reduction system must, at the very least, be off-set by the use of extra equipment elsewhere. Zoom and slow motion are likely to be used chiefly for sports programmes, and each will inevitably cause a loss of vertical resolution on account of the interpolation that must be used. In one case this is used to alter the size of the picture, but, for the other, each individual field must be replicated which requires odd fields being turned into even fields, and vice-versa.

For video tape recording the 8-field PAL sequence is a nuisance, and in theory component coding would be useful. In practice, however, the problem can be

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* In practice a small change of sampling frequency for component coding, equivalent say to the removal of the PAL half-field frequency off-set, is permissible provided an equivalent minimal change of picture geometry can be tolerated.

overcome by permitting a sideways displacement of the picture of one half-cycle of sub-carrier and/or a vertical movement of one line of a field. Obviously this would be unsatisfactory for animation but the 8-field sequence is not a problem in this type of production.

To obtain adequate results for these particular operations where the advantages of component coding are particularly favourable would require very high bit rates. The luminance signal would require to be sampled at least three and probably four times sub-carrier frequency, and each colour-difference signal would require to be sampled at sub-carrier frequency. This gives a total bit rate of nominally 213 Mbit/s without error protection. Although this high bit rate can be achieved, it would not be economic as the basis of a standard for general use. However, it may be required within specialised processing apparatus.

Should a special component-coded signal be used for those purposes where it is particularly advantageous, the choice for the remainder of the system remains open between component and composite coding, but the codes adopted must be compatible with the special component code. Whilst the bit rates involved do affect the decision, and are considered in the next section, the most important question to be considered here is how the choice of component or composite coding will affect the introduction of digital processing within an existing analogue PAL environment.

From the source to the input of the PAL coder the analogue signal is usually in the form of the red, green and blue primaries. It would seem logical either to digitally code the red, green and blue signals independently, or to use luminance and chrominance component coding. The chrominance component could carry the $U + V$ signal on one line and the $U - V$ signal on the next. The use of these sum and difference chrominance signals could very much simplify the realisation of a digital PAL coder. After the PAL coder the analogue signal invariably remains in the PAL encoded form. For this reason it would appear logical to use composite digitally-coded signals in mixers and video tape recorders, and for transmission within the network.

By adopting the system of coding most commensurate with the particular form of the analogue signal, the difficulties of introducing digital equipment into an analogue environment will be minimised. It seems likely that much of the digital hardware could be designed such that it would operate with either composite or component coded signals. This would minimise the difficulties and the cost should a change of philosophy eventually become desirable.
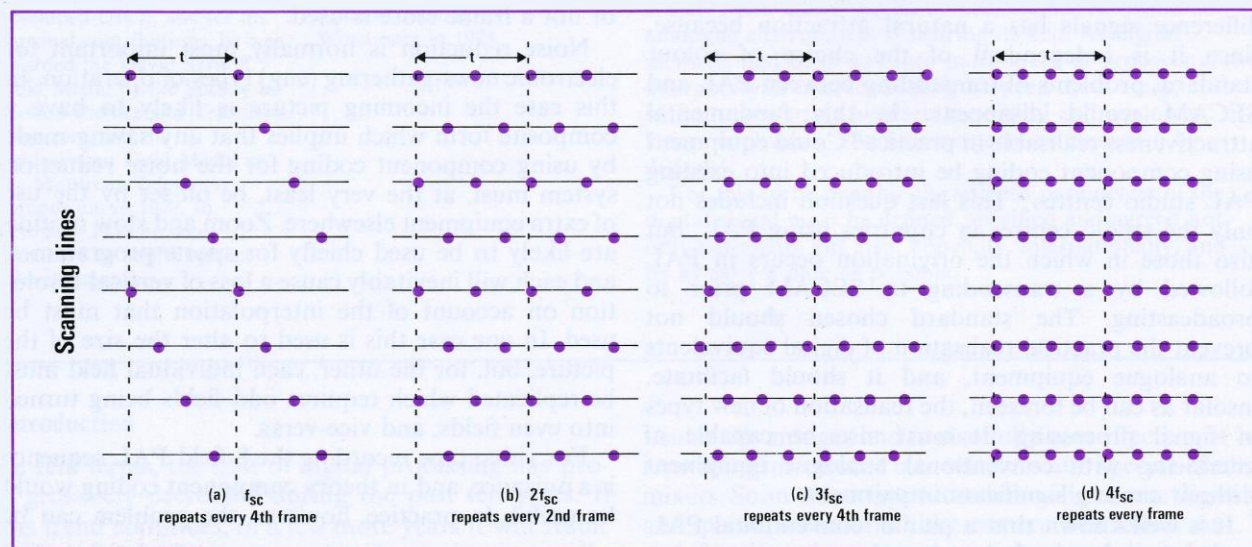
Fig. 1. The diagrams show, for part of a single field, the location of sampling points on a scanned raster for the sampling rates $1f_{sc}$, $2f_{sc}$, $3f_{sc}$ and $4f_{sc}$ respectively. The relationship between sub-carrier and frame frequency in the PAL system results in the patterns repeating at the intervals shown. In each case the time interval, t, equals $1/f_{sc}$. Note the non-orthogonal structure in the case of $3f_{sc}$. This introduces serious difficulties in the digital signal processing, particularly when using two-dimensional image filtering. For these reasons the $3f_{sc}$ sampling rate has lost favour and is not recommended.

Sampling Frequencies and Patterns

If a PAL signal is sampled at sub-carrier frequency the points on the display at which the sampling occurs form a pattern. This pattern repeats once every eight fields and is shown, in part, in Fig. 1(a). (Note: each of the patterns shown in Figs. 1(a)–(d) are for one field only.) It could be appropriate for sampling a single chrominance component, e.g., the U signal, or it may be used for sampling the $U + V$ and $U - V$ signals occurring on alternate lines. In many ways it is undesirable for this pattern to move horizontally from one line to the next and also to move from frame to frame, but it seems better than using a static pattern. This would require a varying interpolation, and the resulting rounding-off errors could cause visible, moving impairments.

Fig. 1(b) shows the pattern of the sampling points when sampling at twice sub-carrier frequency. Due to the sub-carrier having a quarter line-frequency off-set, the sample points on adjacent lines are interleaved. This interleaving enables comb filtering to be used for removing the effects of aliasing which occurs in the spectral region from a frequency equal to twice the sub-carrier minus the highest video frequency, to the highest video frequency. Comb filtering reduces diagonal resolution, and for composite coding also reduces vertical chrominance resolution. Such losses of resolution are rarely visible and do not accumulate with successive passes through similar systems. Signals of this type may be produced from signals sampled at four times the sub-carrier frequency by a digital comb filter, and may be converted back to $4f_{sc}$ signals by a complementary comb filter. The small losses of diagonal luminance and vertical chrominance resolution remain, and in this case the sampling pattern repeats on alternate frames.

Sampling at three times sub-carrier frequency results in the pattern shown in Fig. 1(c) which occupies four frames. This slow repetition rate is a nuisance and the non-orthogonal pattern is especially undesirable because it is asymmetrical about any vertical line. Although use has been made of this pattern it is becoming unpopular for the reasons mentioned above and that, on account of the asymmetry, two-dimensional image processing is almost impossible. It, therefore, cannot be recommended as a standard.

Sampling at four times sub-carrier frequency yields a most desirable arrangement of sampling points producing the orthogonal array shown in Fig. 1(d) which repeats frame by frame. The only undesirable characteristic is the high sampling frequency (17·734475 MHz) resulting in a high bit rate. As already men-
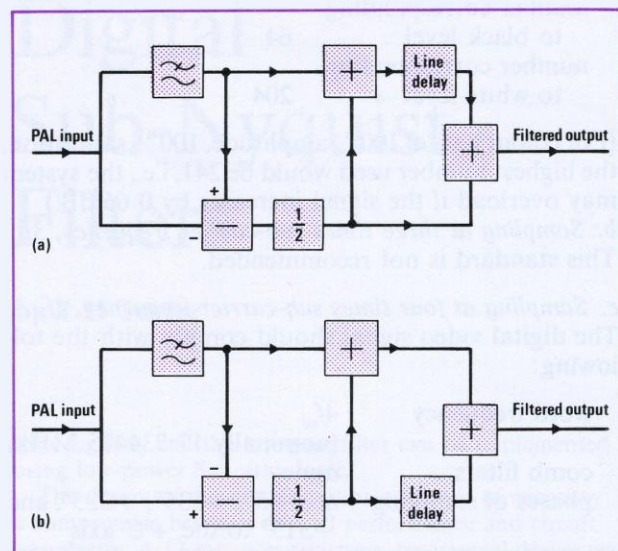
Fig. 2. The block diagrams indicate the preferred form of comb filter to be used at (*a*) the input, and (*b*) the output parts of a system sampled at $2f_{sc}$. The filters may be either of analogue or digital form, but for stability and repeatability the digital approach has considerable advantage. A full description of the digital type of filter can be found in this volume on page 21 et seq.

tioned it is very compatible with $2f_{sc}$ sampling.

Proposed Standards for Composite PAL Coding

a. Sampling at twice sub-carrier frequency, $2f_{sc}$.
Equipment should be designed such that it is capable of accepting as an input, and of providing as an output, the following video signal in digital form:

word frequency	$2f_{sc}$ (nominally 8·8672375 MHz)
comb filters	preferred forms equivalent to those shown in Figs. 2(a) and (b)
phases of sampling	at $+45°$ and $+225°$ to the $+U$ axis
number of bits per word	8*
form of coding	linear pcm

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* If, as seems likely, parallel transmission is to be used for distances below 50 ft, a ninth path will be required for clock information. A further path may be required should error correction or any other information need to pass from one item of equipment to another. For serial transmission, usual for longer distances, the first bit should be the least significant, and successive bits progressively double in significance until the most significant bit is reached. A ninth time slot will be provided which may be used for error control purposes and for ensuring that the clock component can be extracted. The bit frequency shall be $18 \times f_{sc}$ (nominally 79·8051375 MHz).

| number corresponding to black level | 64 |
| number corresponding to white level | 204 |

(For colour bars of 100 % amplitude, 100 % saturation, the highest number used would be 241, i.e., the system may overload if the signal increases by 0·66 dB.)
b. Sampling at three times sub-carrier frequency, $3f_{sc}$.
This standard is not recommended.

c. Sampling at four times sub-carrier frequency, $4f_{sc}$.
The digital video signal should comply with the following:

word frequency	$4f_{sc}$ (nominally 17·734475 MHz)
comb filters	none
phases of sampling	at $+45°$, $+135°$, $+225°$, and $+315°$ to the $+U$ axis

The remaining parameters will be identical with those for the $2f_{sc}$ proposals, except that the bit frequency shall be doubled.

Proposed Standards for Component Coding

As yet it is too soon to be specific about proposals for such standards, but present thinking is centred on using $567\frac{1}{2}$ times line frequency as the sampling rate for the luminance information, and $283\frac{3}{4}$ times line frequency for sampling the chrominance information which latter will be in the form of $U + V$ and $U - V$ signals on alternate lines. These sampling frequencies are virtually identical with twice the sub-carrier frequency, and the sub-carrier frequency respectively.

References
1. Wise, F H, 'The Application of Digital Techniques to Radio Frequency Circuits in Broadcasting Equipment', *IEE Conference Publication*, 1972, 88, pp. 303–309.
2. Baldwin, J L E, 'Sampling Frequencies for Digital Coding of Television Signals', *IBA Technical Review 9*, September 1976.

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Digital Sub-Nyquist Filters

by J H Taylor

Synopsis

Sampling at $4f_{sc}$ rate, full bandwidth, and at $2f_{sc}$ rate, sub-Nyquist, are two proposed standards for use with the PAL signal. The text describes practical, digital, sample-rate changers for converting between these two standards and, at the same time, providing the required comb filtering.

The two basic elements for performing comb filtering in both down and up conversions are a 1-line delay and a low-pass filter. It is possible to operate both of these elements at $2f_{sc}$ rate (8·87 MHz) for converting in either direction and so economising greatly in hardware. The 1-line delay can be realised in the form of an 8×567-bit shift register, and the low-pass filter can be implemented using low-power Schottky logic.

The design of the digital low-pass filter is of necessity a compromise between desired performance and circuit complexity. A 13-tap, non-recursive, transversal design was chosen for the experimental filters described. This resulted in an integrated-circuit package count per filter of about 75, and possibly represents the highest complexity that is practicable, using current technology. Further work is to take place with various low-pass filters, of simplified design, and in assessing their effects on picture quality.

Introduction

The Nyquist sampling theorem states that to digitally encode an analogue signal, the sampling frequency, if unwanted components are to be avoided, must be at least twice that of the highest frequency component contained in the analogue frequency spectrum. Applying this rule to a PAL colour television signal in compliance with the 625-line System I as used in the UK, it will be seen that, without allowing for practical filtering problems, a sampling rate of at least 2×5.5 MHz is theoretically required. However, due to the presence of the colour component in the signal spectrum, it is desirable that the sampling rate should be precisely related to the frequency of the colour sub-carrier, thus preventing the appearance on the display screen of disturbing beat patterns. These considerations have yielded a commonly-used sampling rate of three times the sub-carrier frequency, i.e., $3 \times 4.43 = 13.3$ MHz approx.

It has been demonstrated that due to certain properties of the PAL television signal it is possible, with only slight impairment of the final picture quality,[1] to sample at a rate below that of the Nyquist limit. Most of the energy contained in the spectrum of the luminance signal is clustered around harmonics of the line frequency, and by careful choice of a sub-Nyquist sampling frequency the aliased line harmonic components can be interleaved with the unaliased components. Also, the nature of the PAL signal is such that recovery of the colour information is possible provided the sub-Nyquist sampling frequency is exactly twice the sub-carrier frequency (about 8·87 MHz), and that a particular phase relationship exists between them.

In order to implement this system, commonly known as the $2f_{sc}$ system, it is necessary to employ comb filtering techniques for separating the aliased and unaliased components. Pre-comb filtering, prior to the sub-Nyquist filtering, is also required to improve the performance of the system. Published work[1] provides a ready explanation of the theory of $2f_{sc}$ sampling as applied to PAL System I signals, and describes a system, recently demonstrated, in which the comb filtering is performed on the analogue signal before and after the codec.

Apart from the penalty of high bit rate, a sampling

rate of four times sub-carrier (i.e., 17·74 MHz) has been proposed on the grounds that it provides many advantages, both in signal origination and processing. In terms of the scanning raster, and ignoring the minimal effect of the 25 Hz off-set of the sub-carrier frequency, the spatial relationship between the various sample points forms a regular grid pattern which is repeated during each successive television frame. This property of $4f_{sc}$ sampling considerably simplifies many digital processing requirements.

Either of the two sampling rates mentioned above might be chosen as alternative standards for a PAL System I digital television system. The following describes practical methods of transcoding upwards and downwards between these two sampling rates, at the same time providing the required comb filtering.

Down Conversion ($4f_{sc}$ to $2f_{sc}$)

For a digital signal sampled at $4f_{sc}$ to be prepared for conversion to $2f_{sc}$, it is necessary to ensure that the phases of the sampling pulses relative to the U-axis are precisely maintained. At $2f_{sc}$ sampling rate these are 45° and 225°. It therefore follows that the sampling phases of the $4f_{sc}$ signal are 45°, 135°, 225° and 315°.

It is possible to derive the $2f_{sc}$ signal simply by omitting the 135° and 315° samples, and this is entirely satisfactory in the case of $4f_{sc}$ signals that have been generated electronically, e.g., colour bars (see companion article) and pictures containing no diagonal information. However, the general requirement
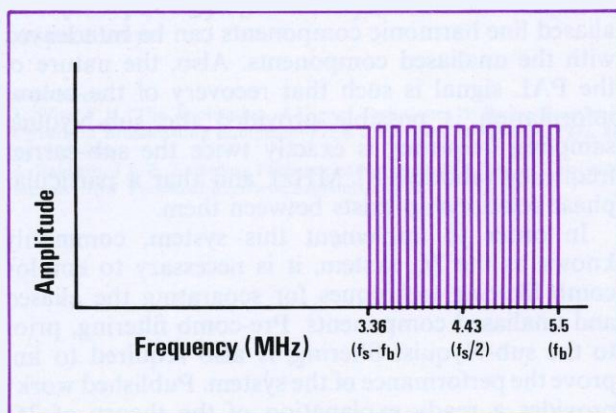

Fig. 1. An idealised one-dimensional amplitude-frequency response for a sub-Nyquist comb filter in which the sampling frequency, f_s, is twice the sub-carrier frequency, i.e., 8·87 MHz. The figures given are approximate for PAL System I in which the video baseband, f_b, is 5·5 MHz. This representation indicates the boundaries for the uncombed low-pass section of the response, and the ideal tooth shape. For clarity the teeth are shown not to scale.
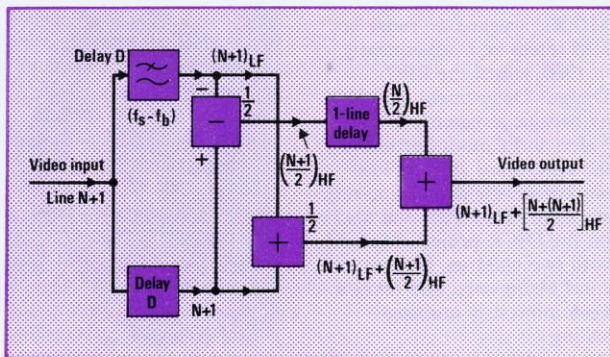

Fig. 2. A form of analogue comb filter using a 1-line delay. The expression for the output signal shows that the low frequencies, defined by the low-pass filter, are derived from a single line only, shown as line $N + 1$ in the diagram. The high frequencies, which are subject to combing and are represented by the tooth shape of Fig. 1, are comprised of information contributed from two consecutive lines denoted by N and $N + 1$.

is for a comb filter to be included at the point of down-conversion.

The requirements of the response of a comb filter for the $2f_{sc}$ sub-Nyquist system are fully described in Ref. (1), while Fig. 1 illustrates the idealised one-dimensional frequency response of such a filter. A similar filter is also required for up-conversion ($2f_{sc}$ to $4f_{sc}$). Figure 2 shows a suggested method for approximating to this response by making use of a 1-line delay in conjunction with the analogue signal. It would be possible to implement this method with a signal in digital format, and with all the functions in the block diagram operating at $4f_{sc}$. The $2f_{sc}$ signal would then be derived by omitting every other sample at the output. However, such a method would be extremely extravagant in terms of hardware, especially the 1-line delay and the low-pass filter.

By a re-arrangement of some of the requirements, it is possible for all the functions of the comb filter to operate at $2f_{sc}$. Figure 4 shows a block diagram of an experimental digital comb filter used for down conversion. But before explaining the method by which the sampling rate is changed reference should be made to Fig. 3. This shows part of the spatial sampling pattern for a PAL signal sampled at $4f_{sc}$, the effect of the 25 Hz off-set again being ignored. The number of cycles of sub-carrier per line is $283\frac{3}{4}$. From this it follows that the 45° sample, for example, advances one sample period to the right on each successive line in every field.

To achieve the comb filtering action, sampling points on any two successive lines must be vertically aligned. By sampling at $4f_{sc}$ this naturally follows,
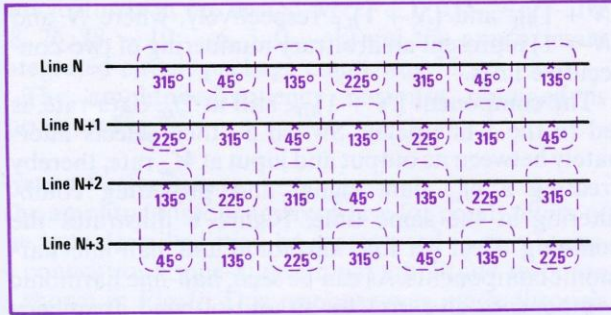
Fig. 3. The diagram here represents the spatial structure of part of a television field in which the $4f_{sc}$ sample points have been indicated. To obtain comb filtering from two successive lines, vertically aligned samples (using a 1-line delay) are required. Because a signal sampled at $2f_{sc}$ requires only the 45° and 225° samples, the 45° sample is processed with the 315° sample on the previous line, and similarly the 225° sample is processed with the previous 135° sample.

but in a $2f_{sc}$ system only the 45° and 225° samples are required. Therefore, as can be seen from Fig. 3, a 45° sample on one line must be processed with the corresponding 135° sample on the previous line; similarly a 225° sample must be processed with the corresponding 315° sample. It follows that the 1-line delay is required to store only the samples at 135° and 315°, the samples at 45° and 225° being routed via the non-delayed path as shown in Fig. 4. Beyond this point, all circuits operate at $2f_{sc}$ data rate. The outputs thus derived from the sum and difference circuits will then comb the frequency spectrum of the input signal, the 'sum' output containing peaks at intervals of line frequency, and the 'difference' output containing peaks at intervals of half-line frequency. The output from the difference path is fed through the low-pass filter, and then combined with that from the sum path to form the output of the complete filter.

The number of samples per line at $2f_{sc}$ rate is $567\frac{1}{2}$ (i.e., $2 \times 283\frac{3}{4}$). The length of digital delay lines is necessarily an integral number of delay periods, and in this case has been chosen to be 567. The alternative of 568 might equally well have been chosen. The extra half sample period enabling addition and subtraction with the undelayed path is achieved by careful latch timing.

Up-Conversion ($2f_{sc}$ to $4f_{sc}$ Sampling)

The post-comb filter, two-dimensional, amplitude-frequency response may be identical with the pre-comb, down-conversion filter already described. However, the requirement for changing the sample rate
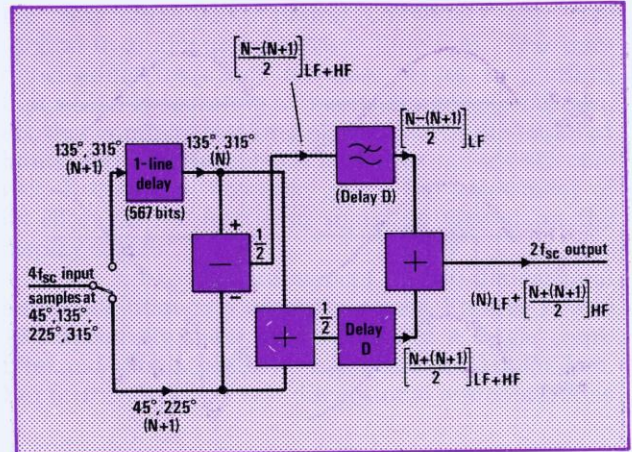

Fig. 4. The comb filter for down conversion as realised in the digital format. Data at $4f_{sc}$ (17·74 Mbit/s) enters the filter and is processed entirely at $2f_{sc}$ (8·87 Mbit/s). The switch shown on the left represents schematically that samples of input data are routed alternately to the two paths as shown. In practice this is achieved by bi-phase $2f_{sc}$ clock pulses. The samples referred to in this and subsequent diagrams are in respect of information contained in two consecutive lines, N and $N + 1$, as indicated.
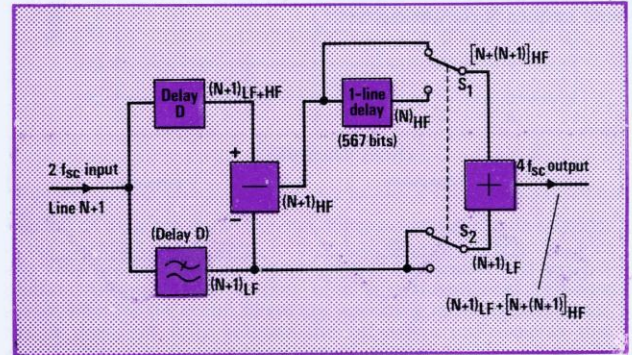

Fig. 5. The full comb filter for up-conversion of data from $2f_{sc}$ to $4f_{sc}$. The data is processed at $2f_{sc}$ rate as far as S_1 and S_2, whereas the final adder operates at $4f_{sc}$. The diagram shows the output from the filter comprising low frequency information from line $N + 1$ only, and the high frequency components from lines N and $N + 1$.

from $2f_{sc}$ to $4f_{sc}$ implies that the configuration of the post-comb filter differs from that of the pre-comb filter. Just as in the case of the pre-comb filter, it is desirable to operate as many functions as possible at $2f_{sc}$ data rate in order to economise on hardware.

Figure 5 shows the block diagram of a practical up-conversion process. Data at $2f_{sc}$ enters a low-pass filter from where it is subtracted from the input signal. This low-pass filter is designed to have a cut-off below the frequency at which aliased components are included. These are indicated on the diagram as
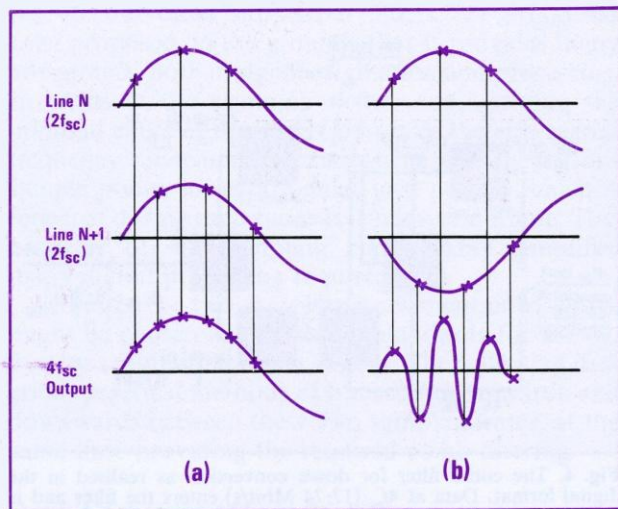
Fig. 6. The action of comb filtering using two consecutive lines, N and $N + 1$, sampled at $2f_{sc}$. The $4f_{sc}$ output is assembled by taking samples alternately from lines N and $N + 1$. The diagram at (a) shows the result derived from two in-phase spectral components of a composite television signal (line harmonic), while that at (b) shows the results from two out-of-phase components (half-line harmonic). It can be seen that this latter produces an out-of-band frequency component at $2f_{sc}$.
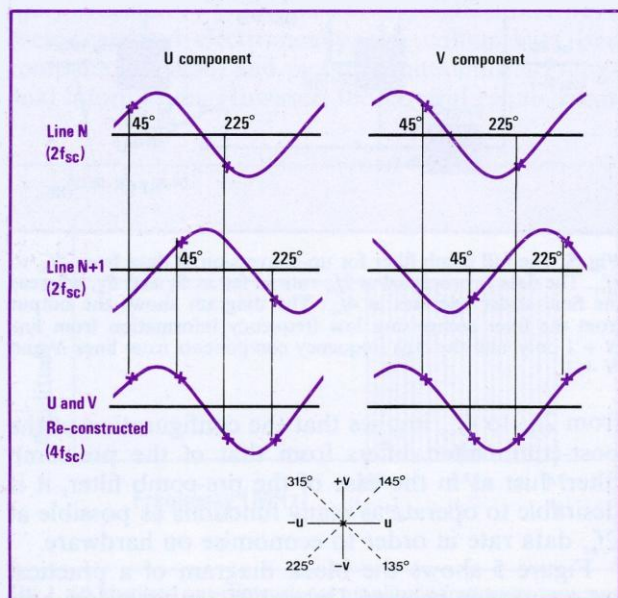

Fig. 7. A simple illustration of how, by using a 567 bit delay element (see Fig. 5), the U and V components of the PAL chrominance signal are reconstructed in the $4f_{sc}$ format from two successive lines of data sampled at $2f_{sc}$. The $2f_{sc}$ samples are taken alternately from lines N and $N + 1$ to form the $4f_{sc}$ data stream.

$(N + 1)_{HF}$ and $(N + 1)_{LF}$ respectively, where N and $(N + 1)$ represent an arbitrary numbering of two consecutive lines.

The component $(N + 1)_{HF}$, still at $2f_{sc}$ data rate, is fed to the 1-line delay. Switch S_1 then selects alternately between its output and input at $4f_{sc}$ rate, thereby creating a $4f_{sc}$ data signal and providing comb-filtering at the same time. Figure 6 illustrates the combing effect on line harmonic and half-line harmonic components. As can be seen, half-line harmonic components give rise to an out-of-band frequency component at $2f_{sc}$.

In addition to providing the required combing, the selection of alternate samples from the input and output of the 567 bit delay enables the chrominance information to be reconstituted as a conventional composite signal sampled at $4f_{sc}$ though, of course, it will have the impairments associated with sub-Nyquist sampling. For a full explanation of the reconstruction of chrominance information reference should be made to Devereux,[1] but a simplified representation is given in Fig. 7. This diagram illustrates the U and V components of two consecutive lines, N and $N + 1$, derived from $2f_{sc}$ sampling, and shows how a combination of the sample values from the two lines yields the corresponding $4f_{sc}$ data stream.

The selection of the appropriate sample value from each line is represented by the action of S_1 in Fig. 5. The low-frequency components of the $2f_{sc}$ signal, containing no aliased products, are routed to S_2 which operates in synchronism with S_1. Each $2f_{sc}$ sample is thus repeated twice, and the result added to the output of S_1. The final adder then provides $4f_{sc}$ data, appropriately comb filtered.

The Low-Pass Filter

The same design of low-pass filter is used in both the pre-comb and post-comb digital filters. The low-pass filter controls the point at which the combing of the amplitude-frequency response commences.

For $2f_{sc}$ sampling, the sampling frequency, f_s, is 8·87 MHz, and if the television signal occupies a baseband, f_b, of 5·5 MHz, alias components can occur between approximately $(f_s - f_b)$ and f_b, i.e., within the range 3·36–5·5 MHz. A practical filter usually has to be a compromise between circuit complexity and performance requirements. For the present experimental filter it was decided that a transversal filter with up to 13 taps would be used. Moreover, as the tap weighting coefficients have a great influence on complexity, the final amplitude-frequency response characteristic was further compromised. The tap weighting coeffici-

ents featured in the design were, $-1, 0, 5, -6, -10, 38, 76, 38, -10, -6, 5, 0, -1$, and the approximate integrated-circuit package count was 75.
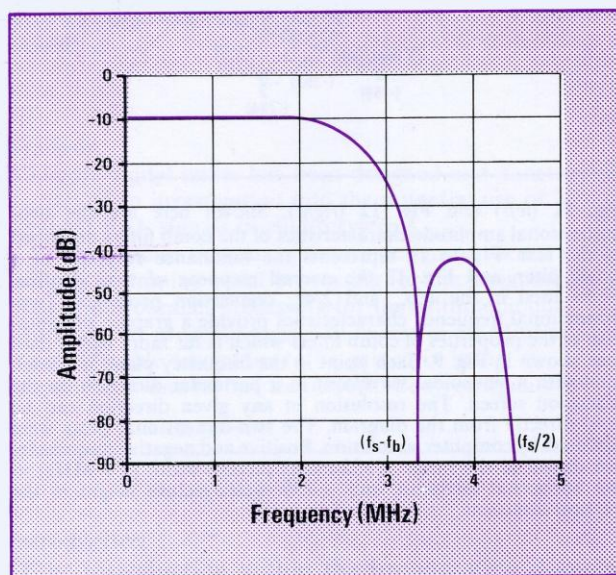
The amplitude-frequency response obtained is shown in Fig. 8.

System Performance

The amplitude-frequency response for comb filters of the type described may be represented in two forms. A conventional one-dimensional amplitude response is shown in Fig. 9. This response was made through a system represented by the block diagram shown in Fig. 10. This includes pre- and post-comb filters in tandem as well as analogue filters associated with

the adc (analogue-to-digital converter) and dac (digital-to-analogue converter) functions. The one-dimensional frequency response serves to confirm that a filter is operating within its design specification, i.e., that combing is taking place between the required limits, and that the cut-offs and attenuations are correct. It does not, however, give a full assessment of its effect on the television signal as viewed on the display screen.

As an aid to this assessment, a two-dimensional amplitude-frequency response is used. Figures 11 and 12 respectively show the response of a single comb filter and of two comb filters (pre and post) in tandem. For further details of two-dimensional representations

Fig. 8. The amplitude-frequency response of a low-pass digital transversal filter having 13 taps, the weighting coefficients of which were $-1, 0, 5, -6, -10, 38, 76, 38, -10, -6, 5, 0, -1$. For sub-Nyquist sampling of the PAL television signal, aliased components extend down to approximately 3·36 MHz.
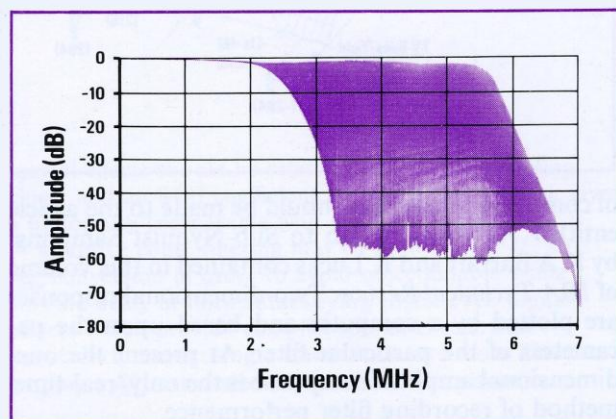
Fig. 9. The diagram shows the one-dimensional amplitude response of the practical system given in the block diagram of Fig. 10. This was obtained by using a swept-frequency synthesiser and an X-Y plotter. Although such a plot of a one-dimensional response shows whether the comb filters are operating in the desired manner, it does not fully indicate the subjective impairments that would appear on the resulting picture display. Figures 11 and 12 show the two-dimensional computer plots which are more useful in this respect.
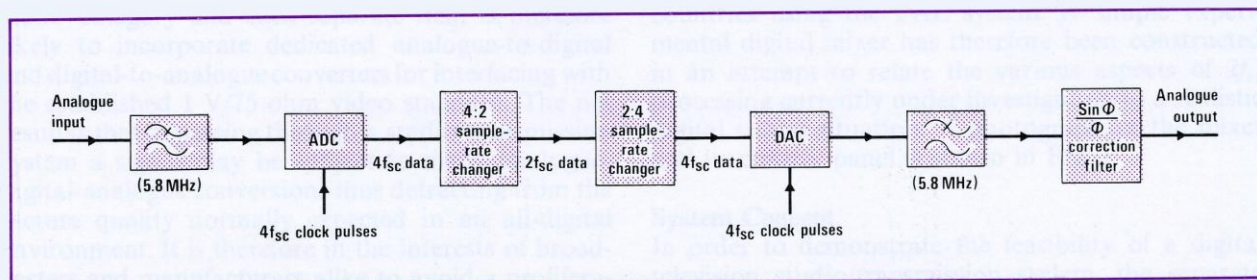
Fig. 10. A block diagram showing an arrangement of pre- and post-comb filters. The sample rate is changed from $4f_{sc}$ to $2f_{sc}$ and back to $4f_{sc}$, producing the overall one-dimensional amplitude-frequency response shown in Fig. 9.
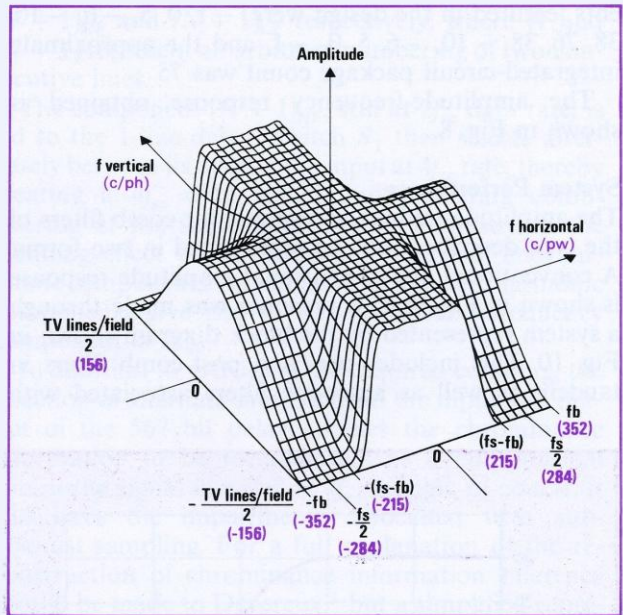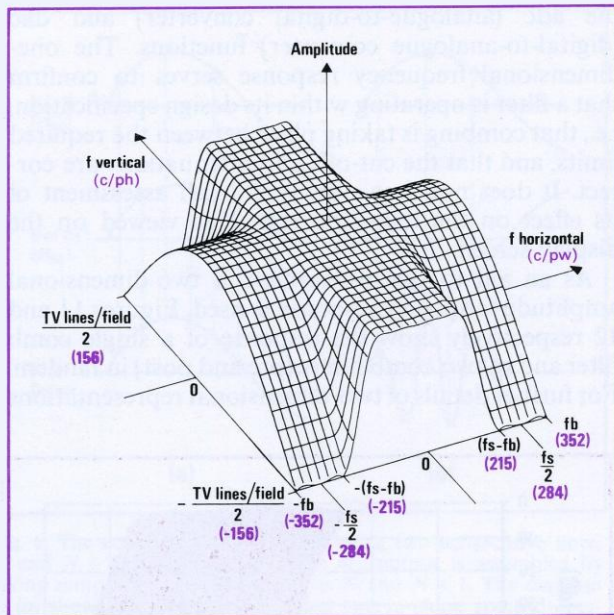
of comb filters reference should be made to the article entitled 'An Introduction to Sub-Nyquist Sampling' by K A Barratt and K Lucas contained in this volume of *IBA Technical Review*. Two-dimensional responses are plotted by a computer and based upon the parameters of the particular filter. At present the one-dimensional amplitude response is the only 'real-time' method of recording filter performance.

Reference
1. Devereux, V G, 'Digital Video: Sub-Nyquist Sampling of PAL Colour Signals', *BBC Research Department Report*, RD 1975/4.

Fig. 11 (*left*) and **Fig. 12** (*right*). Shown here are the two-dimensional amplitude characteristics of the comb filters described in the text. Figure 11 represents the luminance response of a single filter, and Fig. 12 the overall response of two identical filters used in the $4/2f_{sc}$ and $2/4f_{sc}$ conversion processes. Two-dimensional frequency characteristics provide a graphic presentation of the properties of comb filters which is far more useful than that shown in Fig. 9. Each point in the frequency plane is associated with a sinusoidal wavefront in a particular direction on the television screen. The resolution in any given direction can be read directly from the diagram. The two-dimensional plots were obtained by computer simulation. Positive and negative frequencies are shown in the range ±352 cycles/picture width (5·5 MHz) in the horizontal direction, and ±156 cycles/picture height in the vertical direction.

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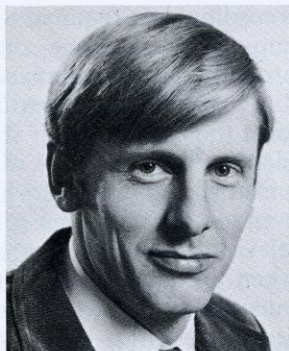
A Digital Television Mixer

by W P Connolly

Synopsis

A simple digital mixer has been designed and constructed as part of an investigation into the potential use of $2f_{sc}$ sampled PAL signals in studio systems.

At the present time an optimum design of such a mixer is likely to employ serial processing for the assignment matrix, but parallel processing for the arithmetic unit. Moreover, at processing rates in excess of 80 Mbit/s there is considerable advantage to be gained in using transmission-line layout techniques. Control signals for special effects should be related to the video sampling clock, and an 8-bit coefficient for fading and mixing gives a uniformity comparable with analogue techniques.

Adaptive comb filtering may be required for $2f_{sc}$ to $4f_{sc}$ conversion in order to avoid objectionable 1-line chroma delay at horizontal keying edges. PAL digital mixers which provide complex special effects, and particularly those using a frame store, are more likely to employ a $4f_{sc}$ orientated component coding system.

Introduction

The introduction within the broadcasting domain of video processing equipment employing digital techniques is increasing progressively. The particular techniques used at present are governed by the needs of convenient implementation and by existing national standards. Most of the equipment falls into the 'stand alone' category and each separate item is therefore likely to incorporate dedicated analogue-to-digital and digital-to-analogue converters for interfacing with the established 1 V/75 ohm video standard. The net result is that in passing through a studio/transmission system a signal may be subject to many analogue-digital-analogue conversions thus detracting from the picture quality normally expected in an all-digital environment. It is therefore in the interests of broadcasters and manufacturers alike to avoid a proliferation of incompatible equipments and standards which would inhibit the eventual transition to completely digital studios and networks with their inherent advantages, and instead to establish national and international standards for the coding of digitised video signals.

Sub-Nyquist sampling of the composite television signal at twice the colour sub-carrier frequency ($2f_{sc}$) is widely accepted as a possible future standard for countries using the PAL system. A simple experimental digital mixer has therefore been constructed in an attempt to relate the various aspects of $2f_{sc}$ processing currently under investigation to a realistic digital studio situation. A photograph of the mixer and its control panel is shown in Fig. 1.

System Concept

In order to demonstrate the feasibility of a digital television studio/transmission system, the separate studies of $2f_{sc}$ sampling, differential pcm bit-rate reduction for link transmission, and digital video

Fig. 1. The 5-input digital mixer and control panel. The signal processing unit is designed for normal 19-inch rack mounting and occupies $3\frac{1}{2}$ inches of rack height. The control panel measures $19 \times 12\frac{1}{4}$ inches and is suitable for either rack or desk mounting.

recording were required to be co-ordinated. The most important tool in the compilation of programmes is probably the vision mixer; therefore, it seemed sensible to combine the various pieces of equipment to form the configuration shown in Fig. 2. Mixers employing digital control of the mixing, fading, wiping and keying functions are already commonplace. Some even incorporate frame stores. These store a complete television picture comprising two fields and are used when producing certain types of special effect. The intention was not to incorporate sophisticated operational features, but to concentrate on the fundamental requirements of cutting, fading, mixing and system timing. To this end a mixer was constructed capable of accepting five source channels, and of assigning outputs to either, or all of three, destination channels and a preview channel. A block diagram is shown in Fig. 3. Input signals selected on the *B* and *C* banks may be horizontally or vertically
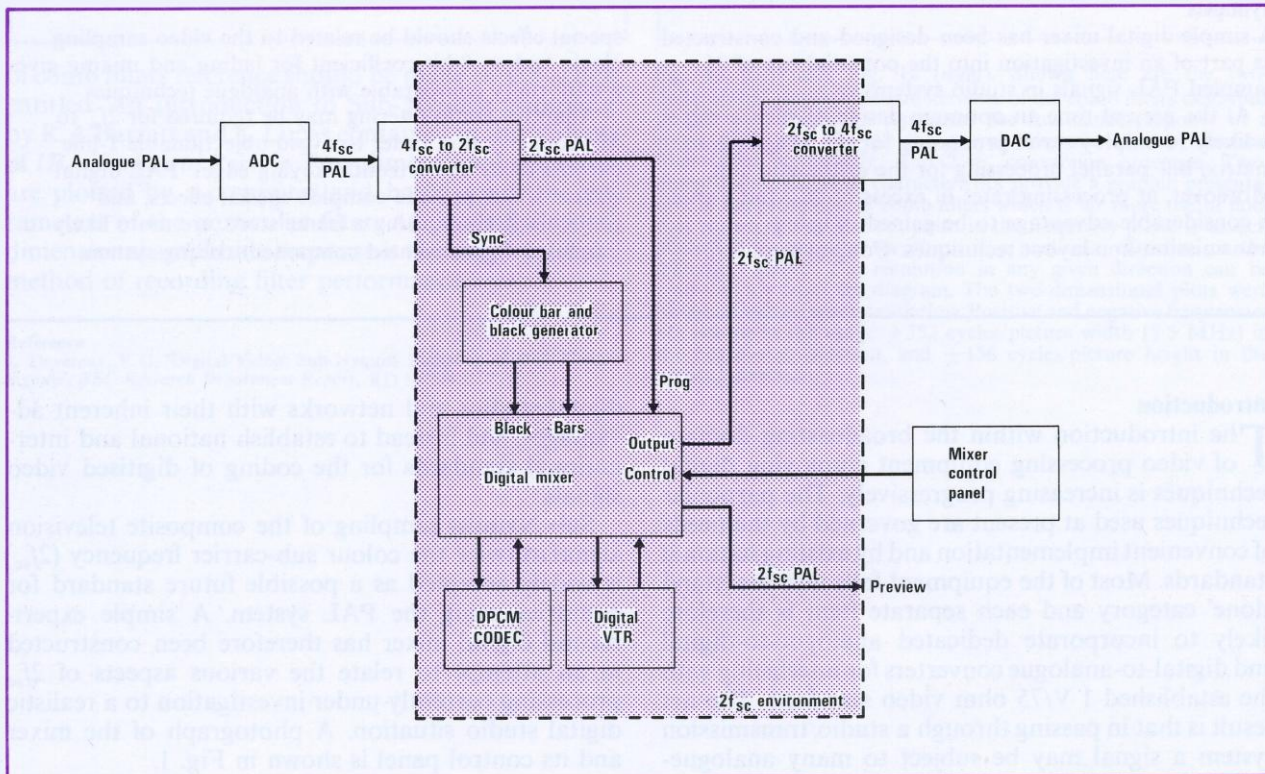
Fig. 2. The elements of a digital television studio. A configuration of programme originating, recording and transmission equipment was assembled capable of demonstrating the feasibility of a completely digital studio/transmission system founded on $2f_{sc}$ sampling. The diagram shows the analogue-to-digital converter in which the composite PAL input signal is sampled at $4f_{sc}$. The sample rate is then converted to $2f_{sc}$ by means of a digital comb filter. Within the $2f_{sc}$ environment, delineated by the dotted line, colour-bar and colour-black signals are digitally generated, and it is with the signals in this form that the processes of mixing, differential coding (to further reduce the bit rate to 34 Mbit/s) and digital recording take place as required. The processed output signal is then digitally filtered back to the $4f_{sc}$ sample rate prior to it entering the digital-to-analogue converter from whence it appears in analogue form.
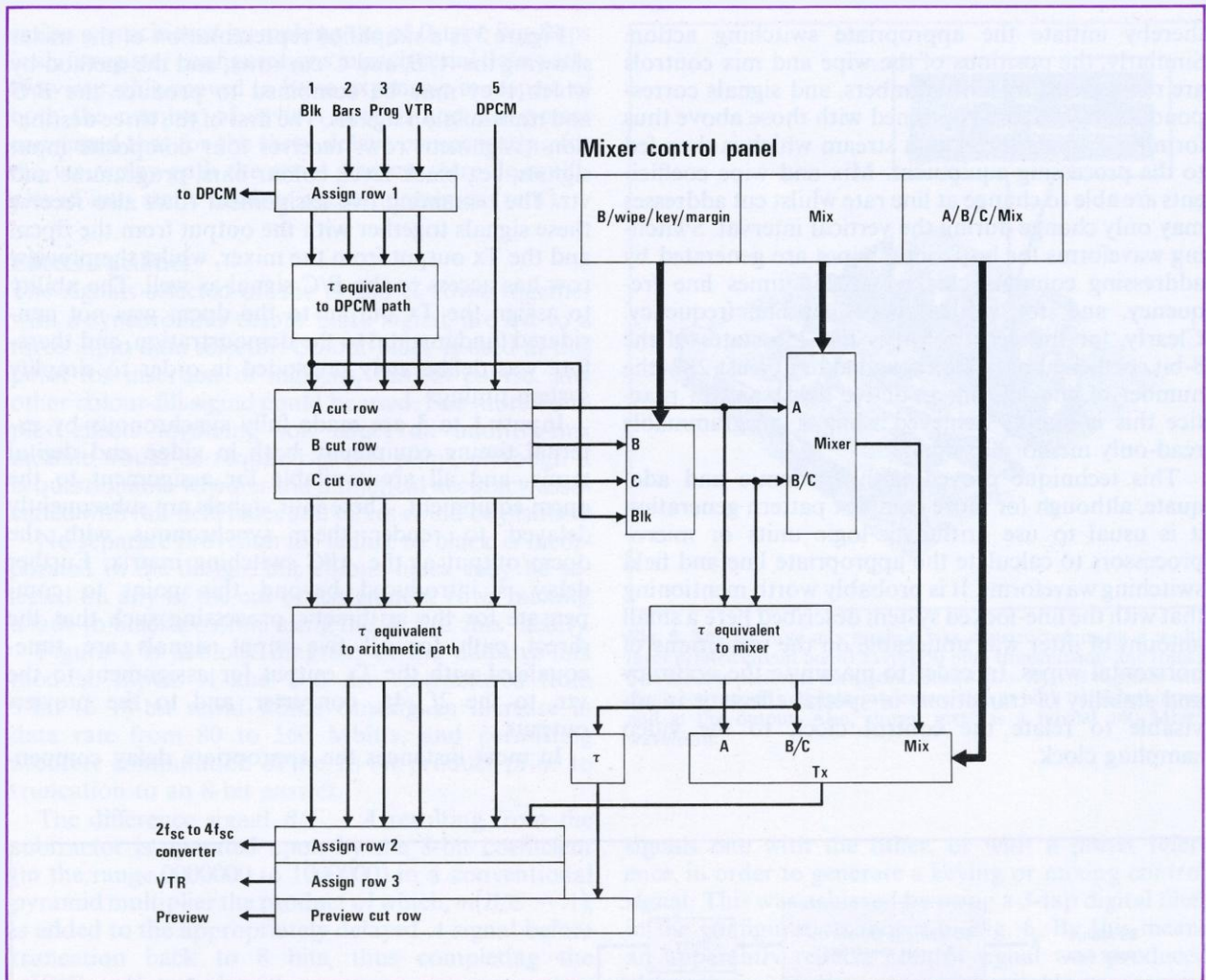
Fig. 3. A block diagram of the $2f_{sc}$ digital mixer which was designed to accept five source channels and to assign outputs to three destination channels. Input channels selected on the *B* and *C* cut rows may be horizontally or vertically 'wiped' with an optional black margin, and the resulting *B/C* signal may be additively mixed to an input signal selected on the *A* cut row. The mixer operates entirely within the $2f_{sc}$ environment, all inputs and outputs being serially coded in linear pcm.

'wiped', and the transition from one to the other may feature an optional black margin. The resulting *B/C* signal may then be additively mixed to the input signal as selected on the *A* bank.

The mixer operates entirely within the $2f_{sc}$ environment, all input and output signals being serially coded in linear pcm.

Recent work has shown that serial processing for video signals is feasible. To test the practicability of this approach it was decided that the mixer should exclusively use serial processing at 80 and 160 Mbit/s.

A mixer is an obvious candidate for serial processing on account of the large number of input/output lines carrying signals which will necessarily be in serial form.

Mixer Design

The pushbutton commands of a particular cut or assignment row are encoded to form signals representing 3-bit binary numbers. These signals are the control inputs of the relevant data selectors in the assignment module of the processing equipment and

thereby initiate the appropriate switching action. Similarly, the positions of the wipe and mix controls are represented by 8-bit numbers, and signals corresponding to these are combined with those above thus forming a single serial data stream which is then fed to the processing equipment. Mix and wipe coefficients are able to change at line rate whilst cut addresses may only change during the vertical interval. Switching waveforms for horizontal wipes are generated by addressing counters clocked at 256 times line frequency, and, for vertical wipes, at line frequency. Clearly, for the vertical wipes the 256 states of the 8-bit coefficient must be expanded to at least 287 (the number of lines during an active field), and in practice this is simply achieved using a programmable read-only memory (prom).

This technique proved both economic and adequate, although for more complex pattern generation it is usual to use arithmetic-logic units or microprocessors to calculate the appropriate line and field switching waveforms. It is probably worth mentioning that with the line-locked system described here a small amount of jitter was noticeable on the transitions of horizontal wipes. In order to maximise the accuracy and stability of transitions in special effects it is advisable to relate the control clock to the video sampling clock.

Figure 3 is a simplified representation of the mixer showing the A, B, and C cut rows, and the method by which they may be combined to produce the B/C and transmission signals. The first of the three destination assignment rows receives four composite input signals, i.e., black level, colour bars, programme and vtr. The remaining two assignment rows also receive these signals together with the output from the dpcm and the Tx output from the mixer, whilst the preview row has access to the B/C signal as well. The ability to assign the Tx output to the dpcm was not considered fundamental to the demonstration, and therefore was deliberately precluded in order to simplify system timing.

Inputs 1 to 4 are made fully synchronous by external timing equipment, both in video and digital terms, and all are available for assignment to the dpcm equipment. These four signals are subsequently delayed to render them synchronous with the dpcm output at the ABC switching matrix. Further delay is introduced beyond this point to compensate for the arithmetic processing such that the direct path for all five input signals are time-equalised with the Tx output for assignment to the vtr, to the $2f_{sc}/4f_{sc}$ converter, and to the preview outputs.

In most instances the appropriate delay compen-
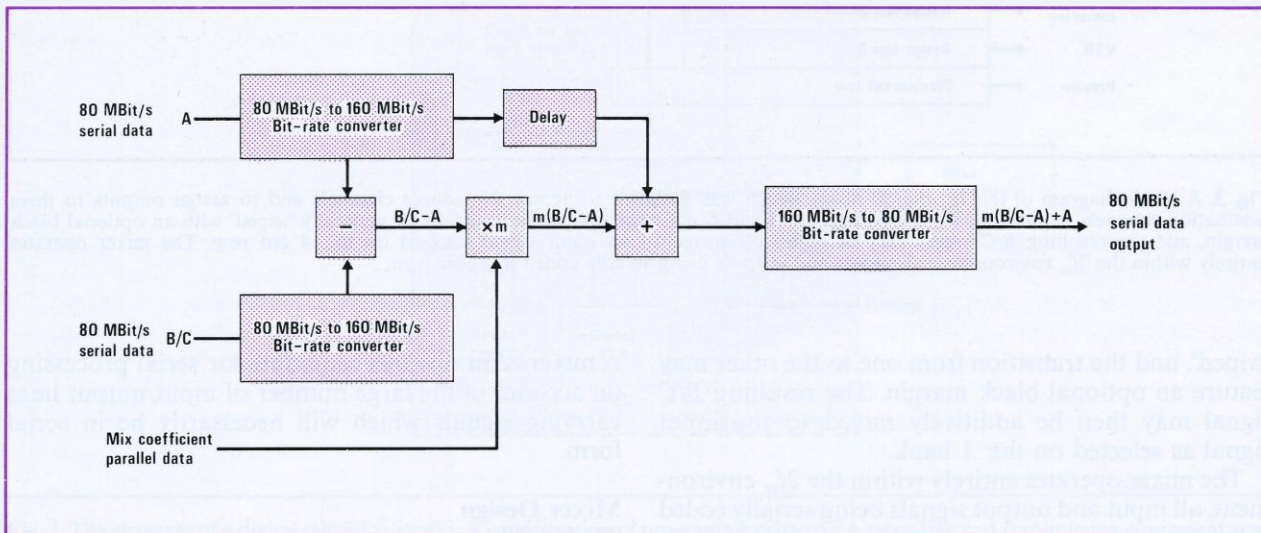
Fig. 4. In the additive mix module, a block diagram of which is shown here, input signals A and B/C are restructured from 9-bit to 18-bit serial words and are subtracted to obtain a difference signal $B/C - A$. This is then operated upon by the mix coefficient, m, ($0 \leq m \leq 1$) in a conventional pyramid multiplier. The output of the multiplier is then added to the appropriately-delayed A signal to complete the $m(B/C - A) + A$ algorithm prior to conversion back to 9-bit serial data at 80 Mbit/s. A photograph of this module is given in Fig. 5.

sation was achieved by making use of *D*-type flip-flops in shift-register configurations in each serial data path. However, this proved to be so expensive in respect of both the number of devices needed and the power consumed that for the three delays compensating for the vtr replay path the serial streams were de-multiplexed into static random-access memories (ram).

Effects Facilities

The signals selected on the *B* and *C* rows, together with a synchronous colour black signal, are fed to a three-input data selector. Colour black is used at this point for insertion of margins, but, of course, any other colour-fill signal could be used. For more complex effects involving soft edges an additive-mix module would be required at this point, although it is questionable whether the numerical accuracy associated with full-field fades and mixes could be justified.

No separate provision for fading to black is incorporated in the design, but colour black may be selected on any of the cut or assignment rows making a fade to black possible using the *A*, *B*/*C* mix facility.

Figure 4 is a block diagram of the additive-mix module. Inputs *A* and *B*/*C* are restructured from 9-bit to 18-bit serial words causing an increase in data rate from 80 to 160 Mbit/s, and permitting accurate computation of the 18-bit product prior to truncation to an 8-bit answer.

The difference signal $B/C - A$ resulting from the subtractor is operated upon by the 8-bit coefficient (in the range 0000000 to 1000000) in a conventional pyramid multiplier the product of which, $m(B/C - A)$, is added to the appropriately delayed *A* signal before truncation back to 8 bits, thus completing the $m(B/C - A) + A$ algorithm.

Most of the additive-mix module operates at a sampling rate of 160 Mbit/s, and considerable attention has been paid to maintaining the quality of signal waveforms. The unit employs a 4-layer circuit board with internal power planes thus permitting all printed tracks to form correctly terminated transmission lines. The multiplier is based on a 1-bit serial adder module capable of working at rates in excess of 300 Mbit/s. A photograph of the mix module is shown in Fig. 5.

In the initial planning of the project it was felt that for demonstration purposes it may be desirable to have the facility for inserting captions into the video signal, and therefore some consideration was given to luminance keying and non-additive mixing (nam) in a $2f_{sc}$ environment. These functions require the continuous comparison of the luminance levels of two
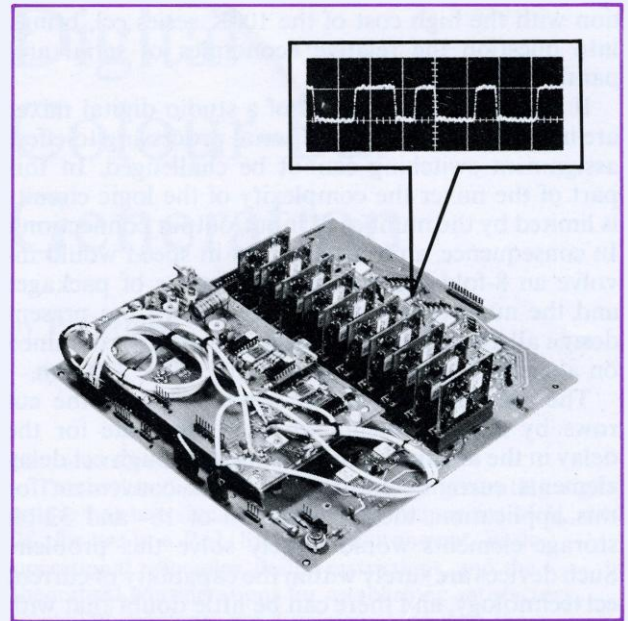
Fig. 5. The additive mix module was constructed using a multilayer printed circuit board having 50 ohm transmission-line tracks. The module performs the algorithm $m(B/C - A) + A$ together with clock conversion and data-rate conversion at both the signal inputs and at the output. Also, shown inset, is a typical 160 Mbit/s waveform.

signals one with the other, or with a preset reference, in order to generate a keying or mixing control signal. This was achieved by using a 3-tap digital filter in the configuration shown in Fig. 6. By this means an apparently reliable control signal was produced although, in the absence of two suitable programme feeds and, more particularly for nam, a second multiplier, a thorough evaluation of the subjective results was not possible.

Design Assessment and Future Considerations

One of the lessons learned in the construction of the mixer was the real advantage to be gained from using the multilayer (transmission line) technique at data rates in excess of 80 Mbit/s. Few problems concerning waveform quality were encountered on the 160 Mbit/s multilayer card, in which 50 ohm transmission-line track is employed throughout. This can be seen in Fig. 5. Conversely, varying degrees of difficulty were encountered with the 80 Mbit/s cards where transmission-line tracks were not used. It would seem that high-speed serial processing would benefit from this technique, but it is expensive and, in combina-

tion with the high cost of the 100K series ecl, brings into question the relative economics of serial and parallel processing.

If the inputs and outputs of a studio digital mixer are in serial form, the use of serial processing to effect assignment switching cannot be challenged. In this part of the mixer the complexity of the logic circuits is limited by the number of input/output connections. In consequence, an 8:1 reduction in speed would involve an 8-fold increase in the number of packages and the number of interconnections. In the present design all assignment switching circuits are contained on a single 3-layer card measuring 150×220 mm.

The assignment rows are separated from the cut rows by delay elements which compensate for the delay in the arithmetic processing. Although ecl delay elements currently available are not convenient for this application, the introduction of 16- and 32-bit storage elements would largely solve this problem. Such devices are surely within the capability of current ecl technology, and there can be little doubt that with ecl devices of the near future there will be a very considerable advantage in using serial processing for the whole of the assignment matrix.
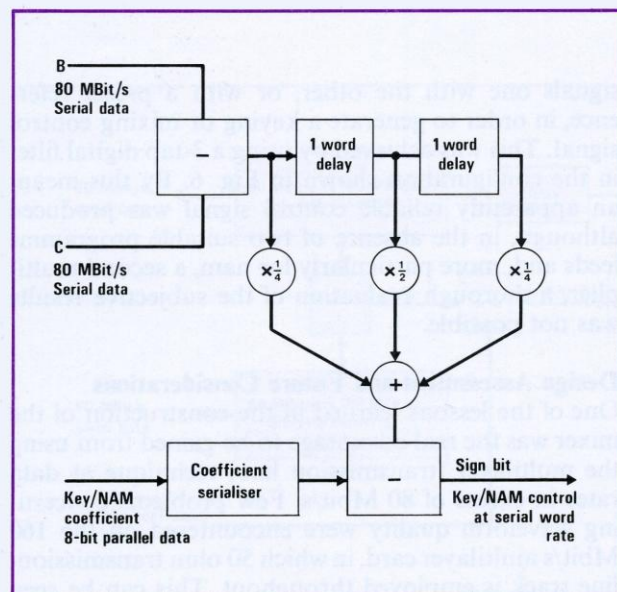
Fig. 6. A block diagram of the luminance key/nam control generator. Luminance keying and non-additive mixing are dependent upon control signals derived by continuously comparing the luminance levels of two signals one with the other, or each against a preset reference. These were obtained using a simple 3-input digital filter as in the configuration shown here.

Even so, it is questionable whether the serial method should be used for the signal processing functions, e.g., additive mixing. In this section of the mixer the complexity of the logic circuits is not limited by the number of inputs and outputs, and, therefore, by making use of the more complex ttl devices currently available, low-speed parallel processing can be of benefit. Moreover, the cost of the integrated circuits, the construction, and the power supply requirements is likely to be lower. The introduction of deserialisers between the assignment matrix and the signal processor is not a serious consideration because an interface would in any case be required for a serial processor in order to effect an increase in word length for arithmetic overflow and subsequent rounding. If parallel interconnections are employed in studios then the parallel method should be used throughout the mixer.

In principle, a mixer working with $2f_{sc}$ sampled composite signals can perform to broadcasting standards all the basic functions of cutting, mixing, fading, wiping, keying, etc. However, subsequent comb filtering, essential to the reconstitution of a colour picture, causes 1-line chroma delay at horizontal keying edges such as are produced in hard-edged special effects. This arises from the interpolation of chrominance information from two lines and may require that the filtering process be made adaptive and that edges be marginally softened. Under normal conditions this 1-line chrominance delay, produced by the $2f_{sc}$ to $4f_{sc}$ comb filter, is balanced by the 1-line luminance delay occurring in the $4f_{sc}$ to $2f_{sc}$ comb filter.

For a simple mixer working with $2f_{sc}$ sampling this small defect is probably insufficient to outweigh the advantages, but the introduction of special effects such as frozen frame, picture rotation, multiple imaging, etc., which make use of a frame store in the signal processor, presents major problems. The processor output must be synchronous with direct signals at the input to the assignment matrix, and at $2f_{sc}$ this requires the introduction of frame-store delays into each of the direct assignment paths. An alternative is to modify the sample values to correct for changes in PAL sub-carrier phase in successive frames. Preferably, this would require a component coding scheme in which corresponding samples in successive frames occupy the same physical position on the screen. The $2f_{sc}$ sample rate cannot satisfy this requirement.

For a digital mixer providing complex special effects it would be preferable to choose a component coding scheme based on the $4f_{sc}$ sampling standard, which can satisfy the requirements described above.

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Digital Colour-Bar Generation

by P Carmen

Synopsis

With the ever-increasing use of digital techniques for television studio systems, the requirement for digitally-generated test signals is becoming greater. The difficulty of doing this is governed to a large extent by the basic television system in question. For instance, in the NTSC system the chrominance signal is provided by a sub-carrier which is line-locked and which alternates in phase from line to line; whereas, in the PAL system, the relationship between sub-carrier and line frequency is complex and only one quadrature component of the chrominance signal changes phase from line to line. Features of this type present complications to the designer and may lead to compromised solutions if unwieldy physical size and high cost are to be avoided.

This article discusses the problems of generating digital colour-bar test signals at sampling frequencies of $4f_{sc}$ and $2f_{sc}$ for use in a PAL System I environment. Basic operational principles, design restrictions, and the theoretical considerations for establishing satisfactory transition envelopes are discussed along with the fundamental method employed for integrating the generator into a demonstration system. In the event, the design proved to be flexible in use and offered the possibility of being readily adaptable to any digital PAL system of compatible characteristics, or even as a self-contained signal source.

Introduction

One of the demonstrations for the Venice meeting of the EBU Technical Committee in April 1977 was centred around a sub-Nyquist system containing a digital mixer. A choice of sources was available each having an independent point of interest to the demonstration, and of these one was a standard test signal, i.e., colour bars, which, it was decided, should be generated digitally as an internal test signal for the system.

In fact, it was considered necessary to provide two sets of bars, one at a sampling frequency of $4f_{sc}$, the other at $2f_{sc}$. In this way a comparison could be made between $2f_{sc}$ originated bars and $4f_{sc}$ bars that had been passed through a $4f_{sc}$ to $2f_{sc}$ conversion process. At first it appeared that two different circuits had to be developed, but it was realised that for signals representing vertical edges only, as with the usual colour-bar format, the $2f_{sc}$ samples could be obtained by simply ignoring every other $4f_{sc}$ sample. With this major feature appreciated, the problem remaining

was to produce, if possible, a single design which could operate in either mode.

Basic Design Concept

Previous experience of digital colour-bar generation for the NTSC system served as a useful basis for examining the problems in producing a PAL generator. The principle evolved was fundamentally the same as for the NTSC generator in that a system of counters, used to define the duration of each bar and each transition, provided addresses to a memory bank. The output of this was then arithmetically processed to give the coded signal.

Fortunately, in handling the PAL signal the chosen sampling rate was $4f_{sc}$. This provided a digital signal which could conveniently be converted into the $2f_{sc}$ sub-Nyquist form. The reason it was so convenient is implicit in the spatial relationship between samples created at this frequency on alternate lines. If, for the moment, the effect of the 25 Hz off-set in the PAL signal is ignored, then the sample positions would be

vertically coherent. However, samples at $3f_{sc}$ rate would be off-set from each other such that the pattern would only repeat after four lines; as will be seen later, the implication of this fact is that much more storage capacity might be required.

In practice, if an examination is made of a signal representing a vertical feature of the display containing chrominance or sub-carrier information, it is found that it cannot be synthesised by identical sample amplitudes repeated from line to line since the 25 Hz off-set has the effect of gradually moving the phase of the sub-carrier relative to the line synchronising signal. Examined in another way, if, in a system sampled at $4f_{sc}$, the four samples per cycle are labelled A, B, C and D, then, in a given field, the PAL $\frac{1}{4}$-line off-set results in the A-samples in any line lying beneath the B-samples of the previous line and above the D-samples of the following line. But, because the 25 Hz off-set adds one cycle of sub-carrier in each complete picture or frame, each sample is displaced from that immediately above it by $225/625 = 0.36$ ns. Thus, an imaginary straight line as near vertical as possible running through $4f_{sc}$ sample points will be found to have a tilt of $0.16°$ which, relative to geometric errors in the display, is negligible. It is worth noting that the only time when this effect can be observed is when a fade is made between $2f_{sc}$ digitally generated bars and converted analogue bars. Even then, a close examination of the green/magenta transition must be made.

Synthesis of sub-carrier by using 'identical sample' values was mentioned above. This is not as difficult as it may sound if the sub-carrier is broken down into its U and V components. Consider the U component; for a constant colour this can be represented by a sinewave delayed by 90° from line to line. Samples at $4f_{sc}$ occurring at 45°, 135°, 225° and 315° all have the same amplitude, and therefore can be represented by a single value but having an appropriate sign in each case. Thus, on following lines the same applies and the one value provides all the necessary amplitude information. A similar argument can be applied to the V component; but, in this case, the signs are inverted on alternate lines to provide the
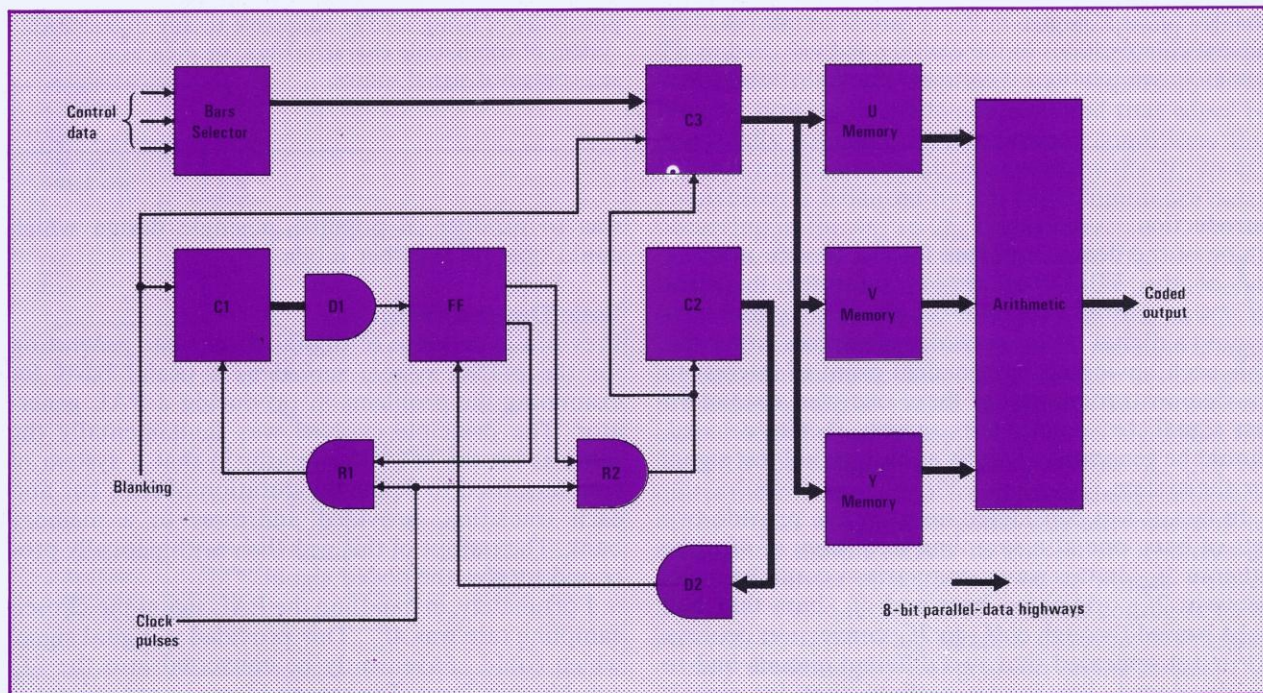
Fig. 1. Block diagram of colour-bar generator. The digital colour-bar signal is obtained by performing arithmetic operations on the outputs of the three memories U, V and Y. Addresses to the memories are provided by a stepping counter, $C3$, which in turn is controlled by the bar-duration counter, $C1$, and the transition-duration counter, $C2$. Counters $C1$ and $C2$ are interconnected by gates $D1$, $D2$, $R1$ and $R2$, and the flip-flop, FF, to determine their start and stop times. Alternative bar signals are obtained by changing the starting address provided by $C3$.

'V switch'. The sum of the U and V components will then automatically contain the appropriate amplitude and sign of each sample.

This principle also applies where the transitions occur between one vertical colour bar and another, but successive samples during a transition are modified in a manner which will be described later. As a result, the vertical edges will slope for the reasons already mentioned. Had this slope not been considered acceptable, its removal would have required an increase in memory capacity of at least 20:1, and a corresponding increase in the arithmetic processing. Incidentally, the $3f_{sc}$ case mentioned earlier, if tackled in the way indicated above, would require either more storage capacity or much more arithmetic processing in order to provide a minimum of two different sample amplitudes for each cycle of sub-carrier.

The considerations described lead to the outline design of the colour-bar generator. Counters, already mentioned, provide addresses to memories organised in ranks containing separate U, V and Y information. In this way, storage is reduced and versatility increased by having independent control of the U and V components.

Circuit Operation
A block diagram of the generator is shown in Fig. 1. Essentially the design uses two counters to address the memory bank. One counter, $C1$, determines the duration of each bar by counting an appropriate number of clock pulses, and its output is decoded by $D1$ to start the second pair of counters, $C2$ and $C3$. The decoded output of $C1$ triggers a flip-flop, FF, feeding clock routing gates, $R1$ and $R2$, so that $C1$ stops and $C2$ and $C3$ start. Counter $C2$ counts nine clock pulses which, when detected by decoder $D2$, are used for resetting the clock-routing flip-flop, FF. The output of $C3$ provides the addresses to the memories and, being fed from the same clock feed as $C2$, the two counters keep in step with each other. The last address present at the end of each count of nine dictates the numbers for the duration of the next bar. During the count the outputs from the memories are changing with every clock pulse, thus providing transitional values. Simple arithmetic functions provide the necessary polarity changes and additions to produce a coded output from the Y, U and V components. (See Appendix 1.)

A range of colour-bar signals is available by taking advantage of memory capacity. These are programmed with three alternative data sets corresponding to 100%, 95% and EBU bars, and are accessed by changing the starting address of the counter, $C3$, thus addressing a different group of memory locations.

Transition Computations
It will be appreciated that colour bars are not produced by a series of step transitions from one luminance level to another, and from one hue to another. Such a method, with zero transition times, would contain wideband spectral components, and the effect of passing these through a limited bandwidth system could produce intolerable ringing on edges. It is therefore desirable that the transition times should be compatible with the system bandwidths for both chrominance and luminance. As shown in *IBA Technical Review 2* the chrominance signal bandwidth is specified as being:

< 3 dB attenuation relative to
           low frequencies at 1·3 MHz,
and
> 20 dB attenuation relative to
           low frequencies at 4.0 MHz;
also, the form of the characteristic should be approximately Gaussian. This definition was used to develop a suitable transition contour for the chrominance components of the colour bars. The luminance envelope was established by considering the normal technique used for assessing system performance, i.e., the use of a $2T$ pulse for which the frequency spectrum, with correct choice of T, is contained within a video bandwidth of 5·5 MHz.

Two stages are involved in establishing the luminance transition,
i to derive an expression for the edge shape so that this can subsequently be included in the computations of sample amplitudes, and
ii to obtain the corresponding frequency spectrum of the edge.

The latter information is required so that the time constant of the transition can be adjusted in order to produce zero frequency components at the sub-carrier frequency, thus preventing the occurrence of interference.

Luminance Edge Shape
The pulse shown in Fig. 2 is mathematically defined as a raised cosine. If the half-amplitude duration is 200 ns then it is more popularly known as a $2T$ pulse. The general form, however, is defined as:

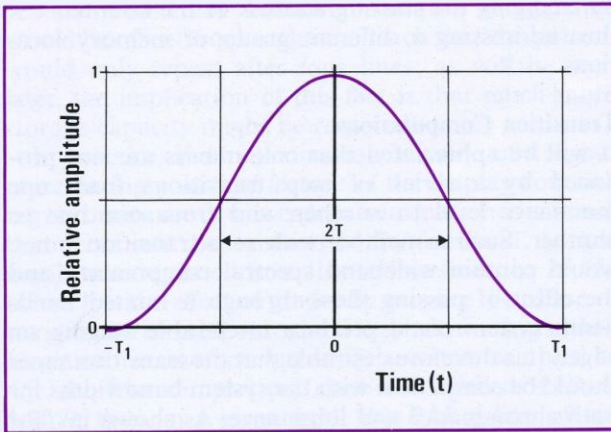$$f(t) = 0.5 \left(1 + \cos \frac{\pi t}{T_1} \right).$$

Fig. 2. The pulse shape shown here is mathematically termed a raised cosine and is given by $f(t) = 0.5(1 + \cos \pi t / T_1)$. Such a pulse, having a half-amplitude duration $(2T)$ of 200 ns and known as a $2T$ pulse, is used for evaluating the performance of television systems. (See Specification of 625-line Video Distortion Measurements, *IBA Technical Review 2*.)

The spectrum of this pulse can be shown to be:

$$F(\omega) = \frac{\pi^2}{\omega(\pi^2 - \omega^2 T_1{}^2)} \cdot \sin \omega T_1.$$

A plot of this is shown in Fig. 3 and an examination of this will reveal that the first spectral zero occurs when $\omega T_1 = 2\pi$. It follows that a raised-cosine pulse will be produced by applying a unit impulse to a filter the transfer function of which is defined by $F(\omega)$. If a unit step is applied to the same filter, then the equivalent time response can be obtained by the reverse transform, thus:

$$F(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} F(\omega) \cdot \frac{1}{j\omega} \cdot e^{j\omega t} \, d\omega,$$

the solution of which is

$$F(t) = 0.5 \left[t + T_1 + \frac{T_1}{\pi} \sin \frac{\pi t}{T_1} \right],$$

and is shown in Fig. 4.

It is worth noting that this can also be obtained by integrating the initial raised-cosine function, $f(t)$. This transition has been chosen for luminance edges in preference to the simpler raised-cosine edge for the following reason. By suitable time-scaling, both transients can be arranged to give a spectral null at sub-carrier frequency (thereby reducing cross-colour effects). However, in the case of the integrated cosine edge, the total energy falling in the chrominance band is lower than that for the raised-cosine transition.

For the integrated cosine edge, the first spectral zero occurs at sub-carrier frequency when T_1 is given by $1/f_{sc}$ (or 225 ns).

Chrominance Transition

For chrominance transitions, the transient spectrum modulates sub-carrier frequency, and the shape of the edge is chosen to achieve a reasonable compromise between the time-response and the cross-luminance energy. This has led to the choice of an integrated Gaussian form. The spectrum of the integrated Gaussian edge is equal to the frequency response of the filter itself:

$$H(f) = \exp\left[-2\pi^2 f^2 \sigma^2\right],$$

where σ = standard deviation.

Now, as the roll-off in frequency response is specified in decibels, the expression can be re-written as:

$$20 \log_{10} H(f) = 20 \log_{10} \left[\exp\left(-2\pi^2 f^2 \sigma^2\right)\right],$$
$$= 20 \log_{10} e\left[-2\pi^2 f^2 \sigma^2\right].$$

Therefore, attenuation, $\alpha = 20 \times 0.4343 \times 2\pi^2 f^2 \sigma^2$. Evaluating σ for $\alpha \leq 3$ dB at $f = 1.3$ MHz, and $\alpha \geq 20$ dB at 4.0 MHz gives 1.01752×10^{-7} and 0.85385×10^{-7} seconds respectively. A value of 0.92853×10^{-7} seconds was taken for use in subsequent calculations.

In the time domain, the transition values may be found by integrating the impulse response of the filter:

$$F(t) = \frac{1}{\sqrt{2\pi\sigma^2}} \int_{-\infty}^{t} \exp\left[-\frac{\tau^2}{2\sigma^2}\right] d\tau,$$

where τ = time
σ = standard deviation (time constant).

The solution to this integral can be found from tables. It is also shown in Fig. 5. By using the value of σ obtained, the time scale can be calibrated; and, if the curve is examined, it will be found that four periods of $1/4 f_{sc}$ taken from the centre point give a value within 0.75% of the initial and final values. Thus, nine transitional values are required to define a chrominance transition to within 0.75% of steady state amplitude during a colour bar. The largest components occur in the V axis for cyan and red, these having a peak-to-peak value of 0.861 V relative to 0.7 V for the white bar. A scaling factor of 128 quanta = 0.7 V gives the definition of 1 quantum on cyan and red bars as 0.63% of bar amplitude. This magnitude of error is beyond the
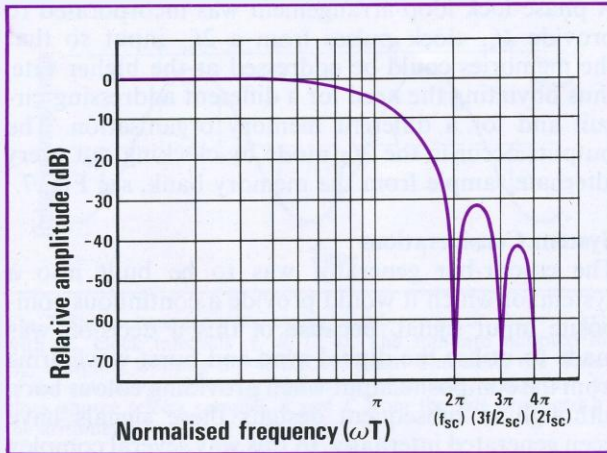
Fig. 3. Frequency spectrum of the raised cosine function $F(\omega) = [\pi^2/\omega(\pi^2 - \omega^2 T_1^2)]\sin \omega T_1$, where $x = \omega T_1$.
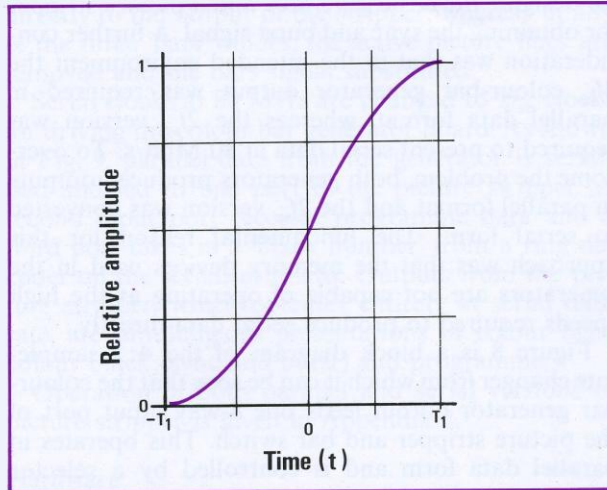

Fig. 4. The sketch shows a normalised transition envelope obtained by integration of the expression for the raised cosine pulse giving $F(t) = 0.5[t + T_1 + (T_1/\pi)\sin (\pi t/T_1)]$.

resolution limits of the system and hence is quite insignificant.

Having established two characteristics, one as an equation for luminance, the other as a set of scaling factors for chrominance, a computer program was written. Operationally, inputs to this program were the definitions of the normal analogue bar amplitudes, scaling factor for luminance reference, bars standards (i.e., 100%, 75%, etc.), lift (appropriate for 95% bars), time constants and black level. Program output consisted of a listing of analogue input levels, scaled 'digital' levels and all transitional values for luminance and chrominance components identified with nominal phase definitions for the $4f_{sc}$ sampling clock frequency. The transitional values were then converted into binary code and used as data in read-only memories. A transition reconstructed from computed values is shown in Fig. 6.

Programming Arrangement

In the previous section, the number of sample values required to reconstruct a chrominance transition was established as nine. Because the bandwidth for chrominance is less than that for luminance, the number of transitional values for luminance must also be within this figure. Hence, the number of memory locations for U, V or Y can be written as:

$$\left(\begin{array}{c}\text{number of sample} \\ \text{values per} \\ \text{transition}\end{array}\right) \times \left(\begin{array}{c}\text{number of transitions,} \\ \text{including black to} \\ \text{white}\end{array}\right)$$
$$+ \text{ (number of bars)}$$

i.e., $9 \times 8 + 8 = 80$.

Two possible memory packages were considered, one having a capacity of 32 eight-bit words, the other having 256 four-bit words. The former was considered initially as this had been used in an established NTSC design, but it became clear that the 256 word device was more suitable. Parallel addressing of two devices could then provide adequate 8-bit wide capability, and the surplus capacity allowed a choice of bar standards. The most straightforward arrangement, subsequently adopted, was to provide two devices for each component Y, U and V using parallel addressing, and utilising the first available 80 memory locations. Bar values and transition values were intermixed as they would appear on the picture, i.e., black/white, white, white/yellow, yellow, etc.

Standards Selection

Both memory packages mentioned above were 16-pin dual-in-line (dil) devices, there being nothing available that was smaller or more economical of storage capability. It was decided, therefore, to make use of the available space for increasing the flexibility of the colour-bar generator by providing a choice of standards. The program was re-run for EBU and 95% bars each requiring a further 80 memory locations. Thus, three sets of numbers were stored requiring 240 of the 256 possible memory locations. Accessing the different bar standards was achieved simply by changing the starting address to the memory bank.
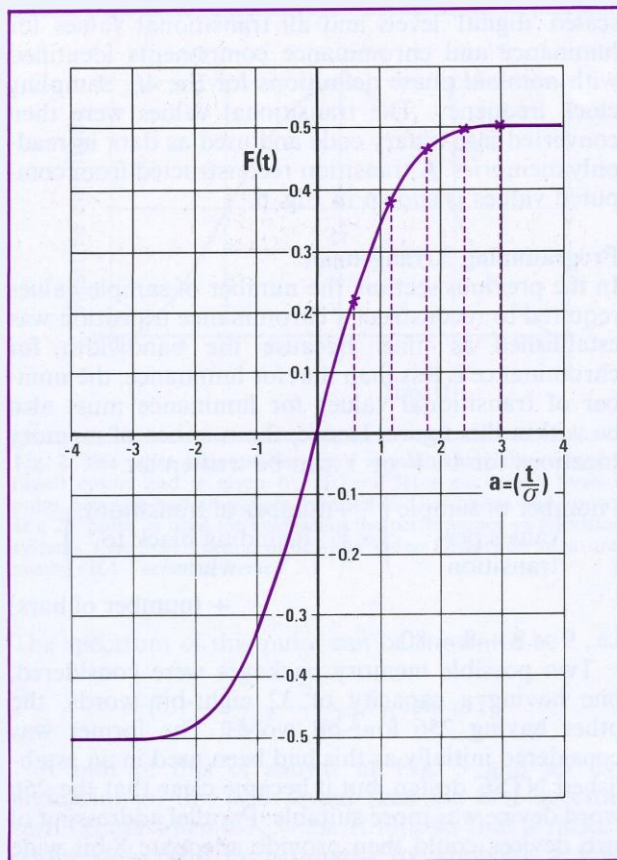
Fig. 5. A plot of the error function $F(t) = (1/\sqrt{2\pi}) \int_{-\infty}^{t/\sigma} e^{-(a^2/2)} \, da$ used to define the limits of transitions in the chrominance signal.

Test Ramp

A rising number, or sawtooth, is an extremely useful signal for use when fault-finding or testing digital systems, and so it was decided to incorporate this as a source in the colour-bar generator. Operation is achieved by clocking a counter from system clocks and enabling the counter when the test ramp is required. Outputs from the counter share a common circuit with the arithmetic processing of the colour-bar generator and therefore the count is arranged to have the appropriate direction, start and finishing points. An option was provided for selecting either a full 256 count, or a count from black level (64) to maximum level (255).

Dual Mode Operation

To enable operation in either a $4f_{sc}$ or $2f_{sc}$ mode it was necessary to include a clock-frequency doubler.

A phase-lock loop arrangement was incorporated to provide $4f_{sc}$ clock pulses from a $2f_{sc}$ input so that the memories could be addressed at the higher rate, thus obviating the need for a different addressing circuit and for a different memory organisation. The outputs occur in the $2f_{sc}$ mode by clocking out every alternate sample from the memory bank, see Fig. 7.

System Considerations

The colour-bar generator was to be built into a system for which it would provide a continuous composite input signal. Because of this a decision was made to utilise the digital sync and burst waveforms from the composite input when providing colour bars, although in subsequent designs these signals have been generated internally. In this way several complex problems were avoided, e.g., storage of sufficient data to describe all the phases and edges of burst present in the PAL signal. This approach left two main areas for design—the basic generator and a picture stripper for obtaining the sync and burst signal. A further consideration was that in the intended environment the $4f_{sc}$ colour-bar generator output was required in parallel data format, whereas the $2f_{sc}$ version was required to present serial data at 80 Mbit/s. To overcome the problem, both generators produced outputs in parallel format and the $2f_{sc}$ version was converted to serial form. The fundamental reason for this approach was that the memory devices used in the generators are not capable of operating at the high speeds required to produce serial data directly.

Figure 8 is a block diagram of the 4:2 sample-rate changer from which it can be seen that the colour-bar generator output feeds one 8-way input port of the picture stripper and bar switch. This operates in parallel data form and is controlled by a selector
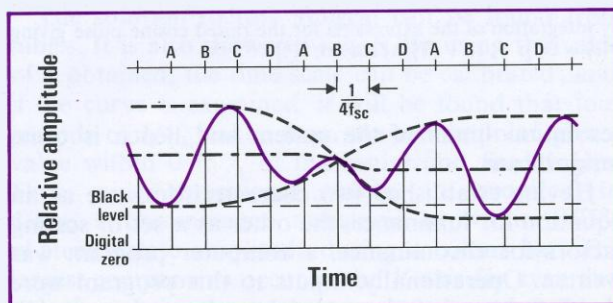

Fig. 6. Reconstruction of a green/magenta transition. Values computed to represent the digital words occurring at the sample instants *A, B, C, D, A, B, C, D, A*, etc., are plotted to give the transition between the two phases of sub-carrier. The envelope of the transition is dotted for clarity.
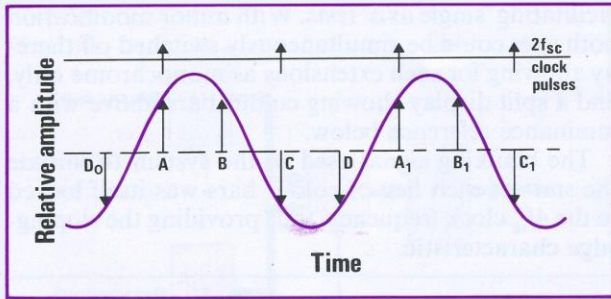
Fig. 7. Shown here are a few cycles of sub-carrier on the $+U$ axis with $4f_{sc}$ clock intervals A, B, C, D. The standard adopted for $2f_{sc}$ sampling is along the 45°, 225° vector. Therefore, the timing of the colour-bar generator has to be organised so that the $2f_{sc}$ clock source selects samples A and C from the sets produced by the arithmetic.
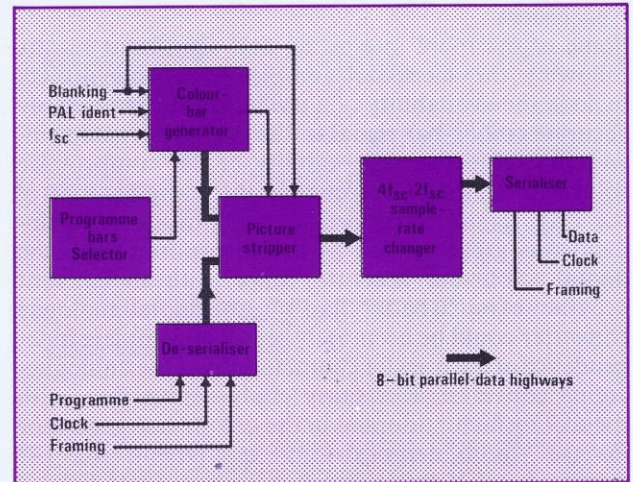

Fig. 8. Basic block diagram of the 4-2 sample-rate changer for use with either programme or 'bars' signals. A programme feed in serial digital form at 160 Mbit/s is de-serialised to parallel data stream at a word frequency of $4f_{sc}$. This signal and the colour-bar data stream, also at the same word frequency, are fed to a picture stripper which, in the 'programme' mode, is transparent. In the 'bars' mode, the active picture period is removed leaving the sync and burst signal to which is added the colour-bar signal. The output of the stripper is then processed by a comb-filter section which changes the word frequency from $4f_{sc}$ to $2f_{sc}$. Finally, the signal is serialised to give an output data rate of 80 Mbit/s.

switch on the front panel of the unit. In the 'programme' mode the programme source is coupled directly to the output of the stripper, whereas in any of the three 'bars' modes, the active picture lines are removed and the bars signal substituted.

Serial clocks at 80 MHz are changed to $2f_{sc}$ clocks for driving the colour bar generator board. As shown in Fig. 9, parallel data from the generator is serialised and fed to one port of the picture stripper. A second input port receives programme data and a third port takes in a 'black number' from a ring decoder on the serialiser board. Outputs from the picture stripper, which operates entirely at serial data rate, are simultaneous presentations of colour bars, colour black (sync and burst) and programme.

Operation of both parallel and serial versions of picture stripper is given in Appendix 2.

Hardware

In terms of size, the complete generator occupies one $9\frac{1}{2}$ in × 6 in double-sided printed circuit board of fairly high packing density. Memory storage consists of six 256-bit 16-pin dil packages, two being necessary for each component U, V or Y, and the remaining circuit functions are implemented with standard '74' series ttl logic devices giving a total package count of 46. Two identical, interchangeable cards were constructed for the demonstration, one operating in each of the modes mentioned. Change from one mode to the other requires only some link changes on the board to accommodate system-timing differences.

Hardware for the picture strippers took the form of two 6 in × 3 in double-sided printed circuit boards, the parallel version containing 16 dil packages compared with 12 dil packages for the serial version.

Fig. 9. A block diagram showing the basic form of the $2f_{sc}$ colour-bar generator and delay equaliser. Serial data at 80 Mbit/s is applied to an input of the picture stripper. In addition a colour-bar signal at $2f_{sc}$ word frequency is serialised and fed to a second input of the stripper, while a third input receives a serial bit stream equivalent to picture black. The three inputs are then processed to provide simultaneous outputs of programme, colour bars and picture black. Delay compensation allowing the picture black signal to be sent to a video tape recorder is incorporated in all data lines as well as clock and framing signals.

Conclusions

Some of the more interesting and significant results following the successful completion of this project include the availability of a useful, stable, predictably accurate colour-bar signal in the digital environment, and the subjective acceptability of the compromise solution reached in overcoming the complications of the PAL signal characteristics. Calibration of equipment was greatly aided by the incorporation of the digital colour-bar generator since, with an inherent accuracy of much better than 1 %, the digital-analogue converter could be accurately set up in complete independence of the transfer characteristic of the analogue-digital converter. It is fairly easily shown that, with the scaling factor adopted in the colour-bar generator for the studio demonstration equipment, i.e., 128 quanta = 0·7 V luminance,* the errors were limited to 1 % in amplitude and 0·3° in phase. These are computed values and ignore any effects subsequent to the digital-analogue process. The presence of two generators within one system aided setting up and fault finding since some elements of the system could be removed (e.g., mixer and delays, dpcm) while the built-in facility for the remaining units was retained. In a real studio situation this would hardly be necessary or justifiable, but in this instance the comparison between $2f_{sc}$ and converted $4f_{sc}$ bars was of interest. As it turned out, theory was corroborated; no differences could be detected between bars generated at $2f_{sc}$ and those converted from $4f_{sc}$.

In ignoring the 25 Hz off-set in order to facilitate the relatively straightforward design and construction of the colour-bar generator, it should be emphasised that the dot pattern is not in any way affected because the relationship between sub-carrier and sync has not been changed.

However, on transitions, a slight difference between the generated bars and the sampled analogue bars was just detectable. This can be explained by the fact that, as the edges are slightly sloping, for reasons already discussed, the transition envelope amplitude is directly related to fixed phases of the U and V components with the result that a degree of regularity occurs. There is some similarity in structure and appearance between these transitions and NTSC colour-bar transitions.

The design of the generator also allows for separate switching of U and V chrominance components, thus facilitating 'single axis' tests. With minor modification both axes could be simultaneously switched off thereby allowing for such extensions as monochrome only, and a split display showing colour bars above with a luminance reference below.

The blanking signal used in the system to initiate the start of each line of colour bars was itself locked to the $4f_{sc}$ clock frequency, thus providing the sloping-edge characteristic.

APPENDIX 1

A square wave at $1/f_{sc}$ rate was used which had a constant phase relationship with the reference burst of the incoming programme signal. By feeding this into a delay line consisting of D-type latches clocked at $4f_{sc}$, the correct phases for the U and V components could be defined since system delays were in integral multiples of $4f_{sc}$. Figure 10 shows the corresponding phase relationship for the U and V signals for the case of two consecutive lines of a field. The $-V$ phase required for every other line is obtained by inverting the phase of the appropriate square wave with the PAL IDENT $\frac{1}{2}$ line-frequency square wave.

If the numbers computed for positive U and V phases of line 1 are stored, then the change of signs for the numbers read out of the memory on subsequent lines will occur automatically, i.e., sample A has a positive U value on line 1 and a negative U value on line 2 though the amplitude is the same. It,
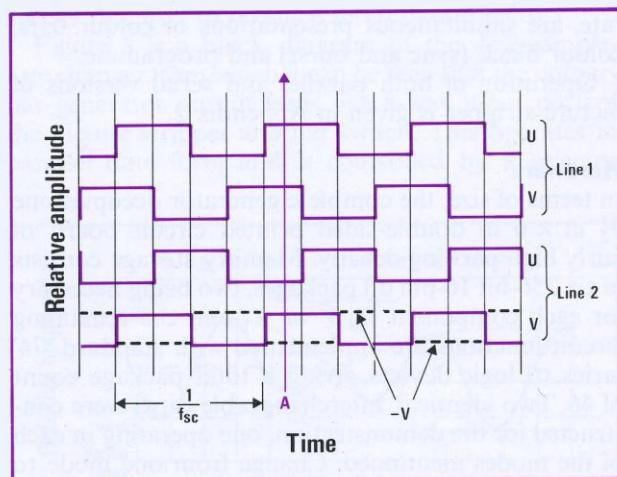
Fig. 10. The figure shows an arbitrary phase of square wave at sub-carrier frequency assigned to the U component on line 1. A square wave defining the V component on $+V$ lines is bracketed with it to indicate relative phase relationship. Line 2 shows that the 'U' square wave has advanced by 270° (due to the $\frac{1}{4}$ line off-set in PAL) while the 'V' square wave, shown dotted, has advanced by 90°.

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* It should be noted that any dynamic range featured as part of a future standard could be accommodated without difficulty (see article entitled 'Proposed Digital Television Standards for 625-Line PAL Signals' on page 16 of this volume).
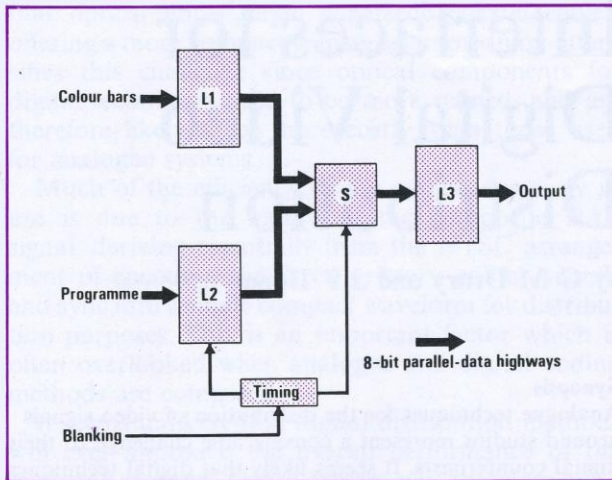
Fig. 11. Block diagram of picture stripper (parallel version). The *D*-type data latches, *L*1 and *L*2, are clocked together to produce data words simultaneously at switch *S*. In the 'bars' mode, the blanking signal selects the programme input data during blanking periods and the colour-bar data during active picture periods. The clocked *D*-type output latch, *L*3, provides accurate timing for subsequent processing.

Fig. 12. Block diagram of picture stripper (serial version). The programme input is buffered and split into two paths. One of these feeds through a delay equaliser to the next stage in the system. The other is applied to switch *S*1. This is controlled by the blanking signal and its output comprises the sync and burst components of the programme feed separated by a data 'number' corresponding to picture black. This 'colour black' signal is buffered and distributed as shown. Switch *S*2, also controlled by the blanking signal, selects colour black during the blanked periods, and colour bars during active picture time.
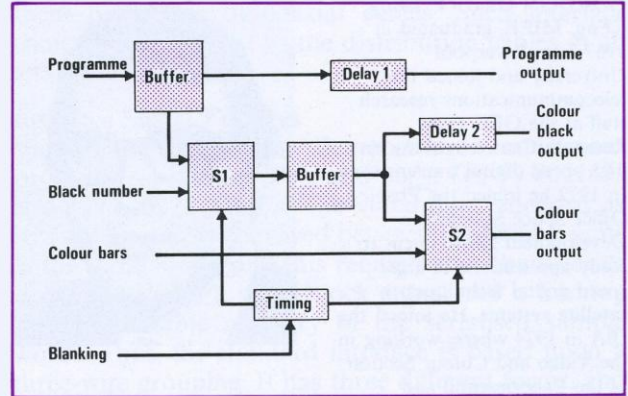
therefore, does not require separate storage. Note that this is true only when the 25 Hz off-set is ignored, otherwise the value would change slightly on every line and so require an inadmissibly large memory.

APPENDIX 2
Picture Stripping

The main purpose of picture stripping is to obtain the sync and burst signal from the programme source so that the active line part of the colour-bar signal can be substituted for the picture content.

The diagram in Fig. 11 shows the fundamental operation of the stripper used for parallel data in the 4-2 sample-rate changer. Operation of the stripper is controlled by the ttl blanking signal which is fed into a timing and gating network from which two modified blanking drives having slightly different pulse durations are obtained. The shorter of the two is used to control the gate latch *L*2 in such a way that the output from the latch is held at black level for a little longer than the active line time but allows the sync and burst components to pass through. The longer blanking pulse holds the switch *S* in favour of the output of *L*2 during sync and burst periods, but

switches over to *L*1 to accept the digital colour-bar signal during the active line time. Thus, switching transients or picture breakthrough cannot occur. The output of the switch is latched via *L*3 to the next system function.

In the serial version of the picture stripper, an outline of which is shown in Fig. 12, the control is again achieved using blanking to select the required part of the line. However, as simultaneous outputs of programme feed, colour black (sync and burst) and colour bars were required in the system, delay compensation and buffering for programme and colour black had to be provided as indicated. Switch *S*1 operates in a similar fashion to the gated latch in the parallel version except that it selects between the programme source and a 'black number' as this was easier to implement with the serial format than by trying to control a single data bit in each word group. Switch *S*2 performs the same function as switch *S* in Fig. 11 and provides an output of serially-coded composite colour bars. Although problems of data timing were greater for the serial picture stripper, its overall complexity relative to the parallel version was not significantly increased.

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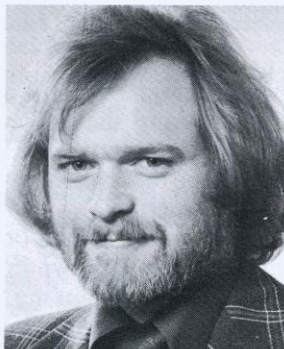
Interfaces for Digital Video Distribution

by G M Drury and J F Dunne

Synopsis

Analogue techniques for the distribution of video signals around studios represent a considerable challenge to their digital counterparts. It seems likely that digital techniques will not meet this fully and that, economically, compensation must be sought in other features of digital studio design.

For the demonstration equipment account had to be taken of some special requirements. Broadly speaking, the problems of studio distribution can be envisaged from three viewpoints. These are local distribution where major processing equipments are close together, distant distribution where these equipments might be separated by distances of up to a few hundred metres, and single-wire distribution in which all the interface information is carried between units on a single wire. This latter arrangement is the most desirable, but may lead to increased costs of distribution compared with analogue methods. All three points of view lead to different requirements in terms of signal formats, coding arrangements and hardware, and these considerations are briefly discussed in the context of the equipment used for the EBU demonstration.

Introduction

The introduction of digital signal processing techniques into broadcasting studios will require solutions to many problems. One of these is the distribution within a studio of digitally encoded sound and video signals as effectively and as economically as possible. This article discusses some of the possible techniques for solving the problems associated with the distribution of digital video signals only. They are certainly not the ultimate solutions, but serve to highlight the system problems and engineering problems associated with this aspect of digital studio design. The equipment designed and constructed for the EBU demonstration provided an opportunity for experiment, and from this some practical knowledge of a working environment has been gained.

Current Techniques

The distribution of composite PAL analogue video signals within modern television studios is performed very efficiently by means of distribution amplifiers and coaxial cables. This arrangement has evolved to a stage where it presents a formidable challenge to its digital counterpart in terms of simplicity, bandwidth, power efficiency, and overall economy. The possibility

that optical fibres might supersede coaxial cables, offering a more compact transmission medium, intensifies this challenge since optical components for digital techniques need to be more refined, and are therefore likely to be more costly, than those used for analogue systems.

Much of the efficiency of the system currently in use is due to the nature of the composite PAL signal, deriving essentially from the NTSC arrangement of encoding the three primary colour signals and sync into a single compact waveform for distribution purposes. This is an important factor which is often overlooked when analogue and digital coding methods are compared.

It seems unlikely that digital distribution methods will ever approach the overall performance of the analogue method. Other factors will therefore need to be considered when choosing a distribution method for an overall digital studio configuration as the possible additional cost, both technical and economic, of digital distribution could be justified only in terms of a saving in some other features of the studio.

The ideal solution in the digital video studio is a single cable connection with simple processing at the terminals. To achieve this, the amount of terminal processing involved, particularly at the receive end, could be considerable and might require the use of custom-designed integrated circuits.

Equipment Configuration

A modern television studio is based on the video mixer, for which the various input picture sources such as cameras, telecine, vtr, slide-scanners, etc., are at different locations in the studio, possibly at distances of a few hundred metres. A similar but less complex arrangement exists for the sound mixer. In either case, the distribution of signals between production and processing areas is performed by groups of tie cables some of which are connected via a patch panel, thereby affording some flexibility of operation. Any conversion to digital operation will probably be required to follow this basic pattern.

The equipment used for the EBU demonstration was arranged to simulate a small scale studio and was based upon a digital mixer the inputs to which were from analogue-to-digital conversion equipment, a digital test-signal generator, digital vtr and a bit-rate reduction unit. These equipments and the complete configuration are described in companion articles. The important feature relevant to this article is that in this equipment the major processing units are discrete and, therefore, that signals are conveyed between

them by means of coaxial cables. Such interface cables are equivalent to the distribution cables in an analogue studio.

Interface Signal Features

Video encoding in the demonstration equipment produces bit rates of $18f_{sc}$ and $36f_{sc}$, i.e., 79·8051375 and 159·610275 Mbit/s. The video data, encoded as straight binary, is conveyed between processing units in the serial mode, and this requires that the receiver is provided with a serial clock signal and a framing signal to enable recovery of the serialised sample words. Thus, the standard interface is based upon a three-wire grouping. It has three different forms, and two possible bit rates;

i a 'local' interface between processing units when these are closely situated, e.g., of less than two metres separation. This is the form mainly used for the demonstration equipment since most of the units were close together. It operates at both $18f_{sc}$ and $36f_{sc}$, corresponding to $2f_{sc}$ and $4f_{sc}$ video sampling rate,

ii a 'distant' interface when, with long cable runs, account must be taken of cable attenuation, induced hum and interference. This operates at $18f_{sc}$ only, and is thus confined to $2f_{sc}$ sub-Nyquist processing, and

iii a 'single-wire' interface operating at $18f_{sc}$ to demonstrate the feasibility of encoding the digital data, clock and framing signals into a single waveform requiring only one cable for transmission.

For various reasons not relevant here, most of the component units employ parallel processing which requires serialisers and de-serialisers to convert to and from the serial mode for transmission between units.

The three cables used in both the 'local' and 'distant' interfaces must have identical delay for the arrangement to function correctly. In practice the delays are so controlled as to be within 1 ns of each other. With the single-wire interface this need does not arise.

Local Interfaces

Local interfaces are used for conveying the serialised data, clock and framing signals between the major processing units. Each major processing unit contains a serialiser/de-serialiser pair which converts between the parallel and serial modes for transmission between units.

Some interfaces connect between processing units operating at $4f_{sc}$ parallel data rate, others at $2f_{sc}$ parallel data rate. Most of the demonstration equipment uses the latter rate; only connections from the

analogue-to-digital converter (adc) and the digital-to-analogue converter (dac), to the '4 to 2' and '2 to 4' filter units respectively, use the $4f_{sc}$ rate. The serial-bit format and timing specifications are illustrated in Fig. 1.

Because the interconnections involved in local interfaces are no more than a few metres in length, no special coding arrangements need be employed and the signal format is retained throughout, i.e., straight binary encoded data and emitter coupled logic (ecl) voltage levels. All cable impedances for this type of interface are 50 ohm. This helps to identify a local interface from a distant one for which 75 ohm cable is used. The choice of impedance was not significant other than for purposes of convenience and standardisation, both within and between units.

The theoretical limit on the distance between units is set mainly by cable attenuation. As cable length is increased, the higher-frequency components become attenuated quite rapidly. Although a local interface could be made to function over distances up to 35 m, in practice its reliability depends upon the ability of the receiver circuits to function on impaired input signals. One of the defining features of a local interface is that no special arrangements are needed to accommodate such factors as long cable runs. Thus, it is advisable that the cable runs employed be kept to an absolute minimum. In the demonstration equipment, the maximum length of interconnection cable was less than 5 m.

Distant Interfaces

A distant interface is required where separation between the units to be interconnected is greater than, say, 20 m. The particular problems arising from the use of these longer cable runs are that dc coupling is
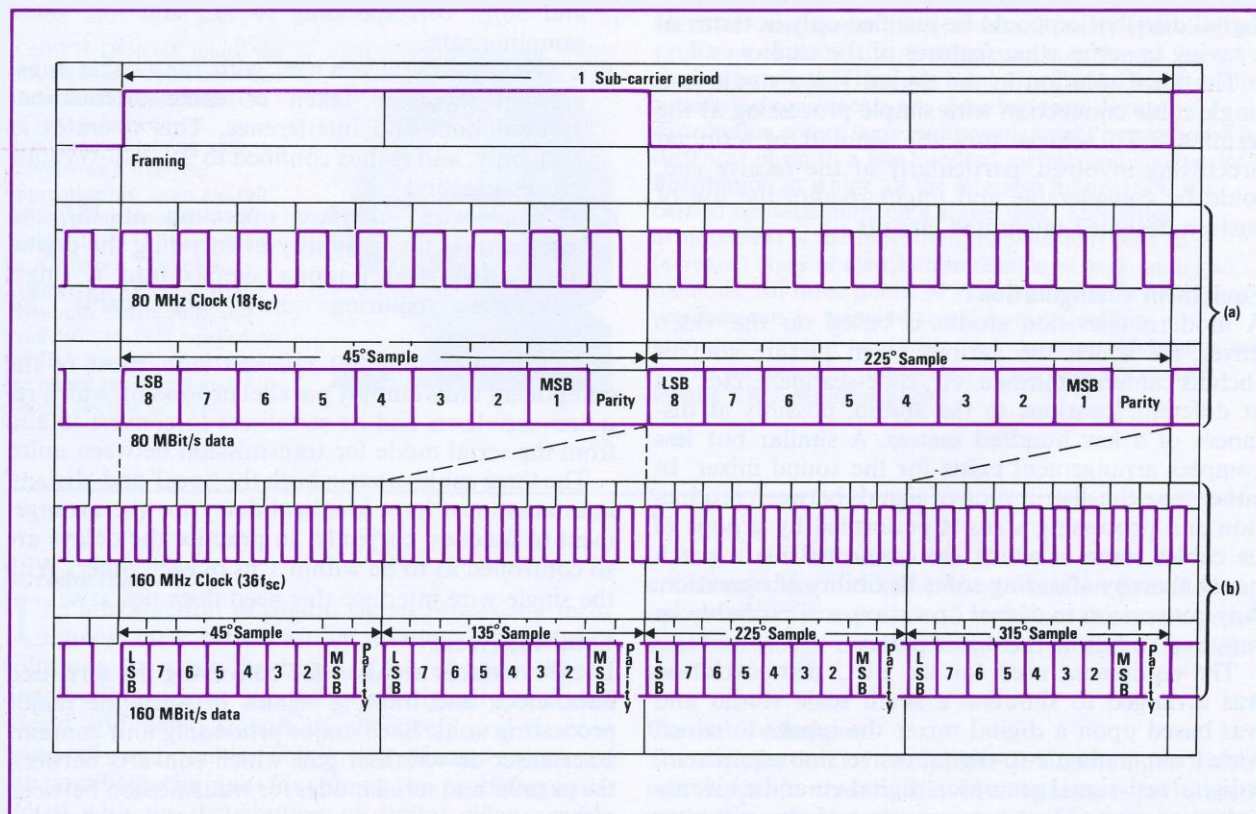
Fig. 1. Serial data structure and timing diagram. The standardised time relationship between the data, clock and framing signals as they appear at the interfaces between processing units is shown. Note that the data bits are arranged in increasing order of significance with the least significant bit (LSB) first and the most significant bit (MSB) last. In addition to the eight data bits there is a ninth time slot which may, for example, be used for parity. Note also that the $18f_{sc}$ interface handles the 45° and 225° samples only, whilst the $36f_{sc}$ interface handles all four samples in the sub-carrier period.
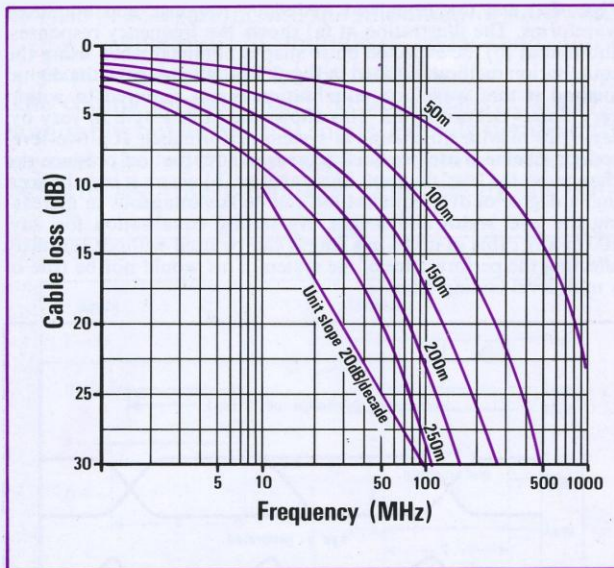
Fig. 2. Loss/frequency response for a typical 75 ohm distribution cable. Cable attenuation against both frequency and cable length is shown. At the clock rates for both interfaces (approximately 80 and 160 MHz) considerable attenuation can occur. The attenuation of the cable is approximately proportional to the square root of frequency in MHz, and, at any fixed frequency, is directly proportional to length. These curves were obtained by arranging a curve to fit the manufacturer's published data and then using the resulting expression to calculate the attenuation. This expression is:

$$\text{Loss (dB/100 m)} = 0 \cdot 151 + 0 \cdot 991\sqrt{f} + 0 \cdot 0151 f \text{ decibels,}$$

where f = frequency in MHz.

less suitable than it is with short runs, and that the equalisation for attenuation/frequency response cannot be ignored.

For a distant interface, the standard arrangement of data, clock and framing signals on three separate cable runs is maintained. The clock and framing signals have no significant low-frequency energy below about 100 kHz. Therefore, for these signals, a change from dc to ac coupling presents no great difficulty. However, the data signal has spectral properties summarised as follows:

a an overall power distribution law against frequency of $[(\sin x)/x]^2$,

b pattern-dependent low-frequency energy at a relatively high mean-power level, and

c a spectral null at the clock frequency.

These properties arise from the non-return to zero (nrz) binary coding format used for the data.

The features of relevance here are the first two, and the use of ac coupling of the straight binary-digit streams that could cause difficulties and eventually

lead to digit errors unless appropriate design precautions are taken. One such design change is to re-code the video information from the straight binary format into one which has an inherently low level of low-frequency energy below, say, 100 kHz. Many methods of achieving this transformation are available; the one used for the present equipment is discussed further in the following section.
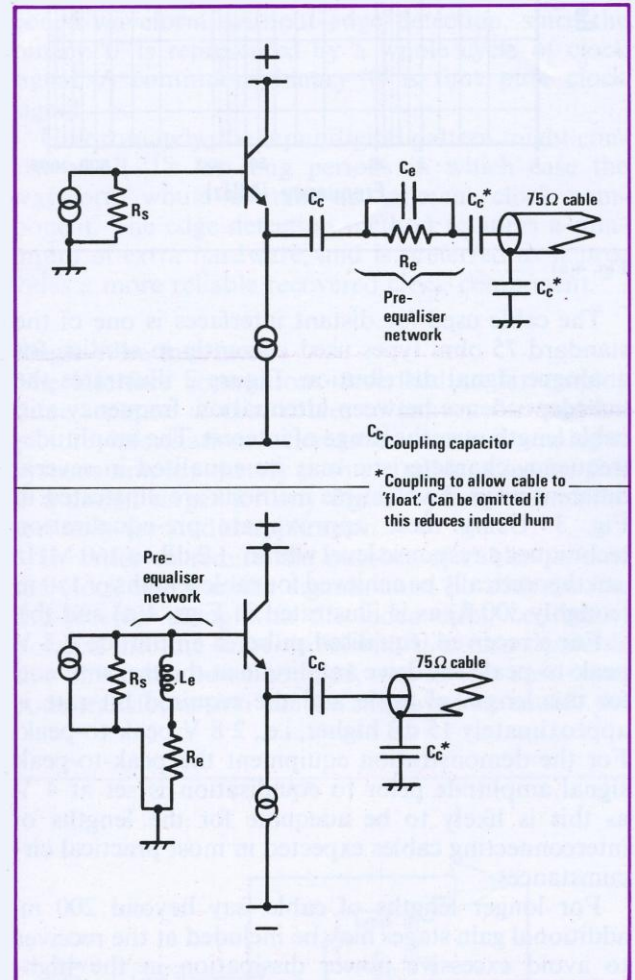
Fig. 3. Simple cable-pre-equalisers. The equivalent circuits shown represent two possible simple methods of applying pre-equalisation to a signal to be transmitted by means of the cable. Each of these arrangements allows, at the distant end, immediate use of the signal without further equalisation. They afford the further advantage (deriving from the digital coding rather than from circuit arrangements) that, once set up, the equalisation of any given cable needs no further adjustment during the life of the system. Moreover, it is tolerant of fairly significant changes in cable characteristics, drift in circuits, etc.
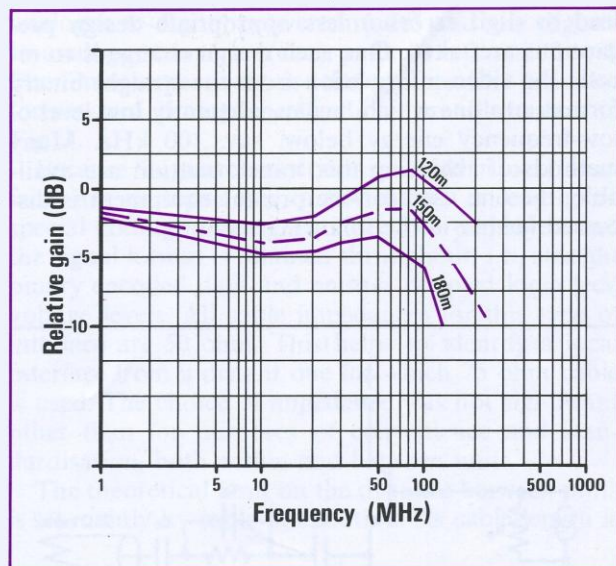
Fig. 4(a)

Figs. 4(a) and (b). Equalised frequency response and equalised waveforms. The illustration at (a) shows the frequency responses, and that at (b) the expected pulse shapes obtained when using the equalisation method outlined in Fig. 3. The advantage of the digital method is that waveform equalisation is not required to within very closely defined limits. This allows the cable length to vary by, say, ± 10 m with no change in system performance. If a two-level coding scheme is used various degrees of equalisation produce the changes in the 'eye' diagram illustrated at (b) where it may be seen that a degree of over-equalisation can be advantageous in increasing the 'eye' width and height. By setting equalisation for, say, 100 m any cable of up to this length can be used without seriously affecting the performance of the system. This would not be true of a multilevel coding scheme.

Fig. 4(b)

The cable used for distant interfaces is one of the standard 75 ohm types used currently in studios for analogue signal distribution. Figure 2 illustrates the interdependence between attenuation, frequency and cable length over the range of interest. The amplitude-frequency characteristic may be equalised in several different ways. Two simple methods are illustrated in Fig. 3. Using these approximate pre-equalisation techniques a response level within ± 2 dB to 160 MHz can theoretically be achieved for cable lengths of 150 m (roughly 500 ft) as is illustrated in Figs. 4(a) and (b).

For a received, equalised pulse of amplitude 0·5 V peak-to-peak, the drive amplitude at the transmit end for this length of cable and the required bit rate is approximately 15 dB higher, i.e., 2·8 V peak-to-peak. For the demonstration equipment the peak-to-peak signal amplitude prior to equalisation is set at 4 V as this is likely to be adequate for the lengths of interconnecting cables expected in most practical circumstances.

For longer lengths of cable, say beyond 200 m, additional gain stages may be included at the receiver to avoid excessive power dissipation in the high-power output stages of the transmitter. Gain stages such as these need not provide any spectral shaping although this may be useful. Hence, simple designs, possibly incorporating an integrated circuit, may be used.

Code Conversion

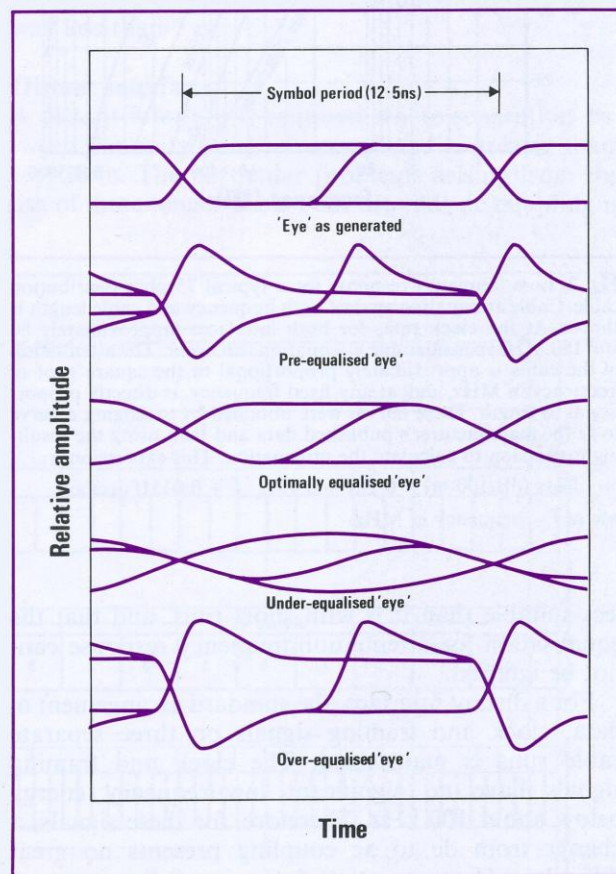The need for code transformation has already been explained. The basic requirements of the new code format are,

1. inherent lack of low-frequency spectral components,
2. the ability to be processed easily for providing clock component, i.e., code must be easily self-clocked. (This is more important for the one-line serialiser which uses the same coding format),
3. controlled spectral properties, especially at high frequencies, and

4. ease of realisation, and the minimum of hardware, e.g., avoidance of multilevel codes and those requiring an increased bit rate.

The coding format chosen satisfies items 1, 2 and 4 admirably, but its spectral energy extends significantly beyond the equivalent digit repetition frequency. Nevertheless, the advantages afforded by the simplicity of this code are considered to outweigh this single disadvantage for the present application.

The basic coding and decoding arrangements are illustrated in Figs. 5(a) and (b) and Figs. 6(a) and (b) respectively. The encoding ensures that a negative-going edge occurs at least once every two symbol periods regardless of input digital patterns. Therefore, by detecting the negative-going edges in the encoded waveform, timing information may be extracted, if required, to enable 'self-clocking' of the code. A clock component could be extracted directly from the encoded waveform, without edge detection, since the binary '0' is represented by a whole cycle of clock signal. A continuous binary '0' is thus pure clock signal.

Unfortunately, the input digital pattern might consist of all '1's for long periods in which case the waveform would contain no inherent clock component. The edge detection method requires a minimum of extra hardware, and is preferred as it provides a more reliable recovered clock component.

Single-Wire Interfaces

The interface connections described so far employ three separate coaxial cables to convey the encoded picture information, whereas the current analogue systems employ merely a single cable. Clearly, this is a design target for any digital system considered seriously for studio application.

In order to reduce the number of cables to one, the clock and framing signals must be conveyed to the receiver along with the video data, i.e., incorporated into the same waveform. It is also desirable that this cable be ac coupled, and in this case some code transformation can be employed to enable the
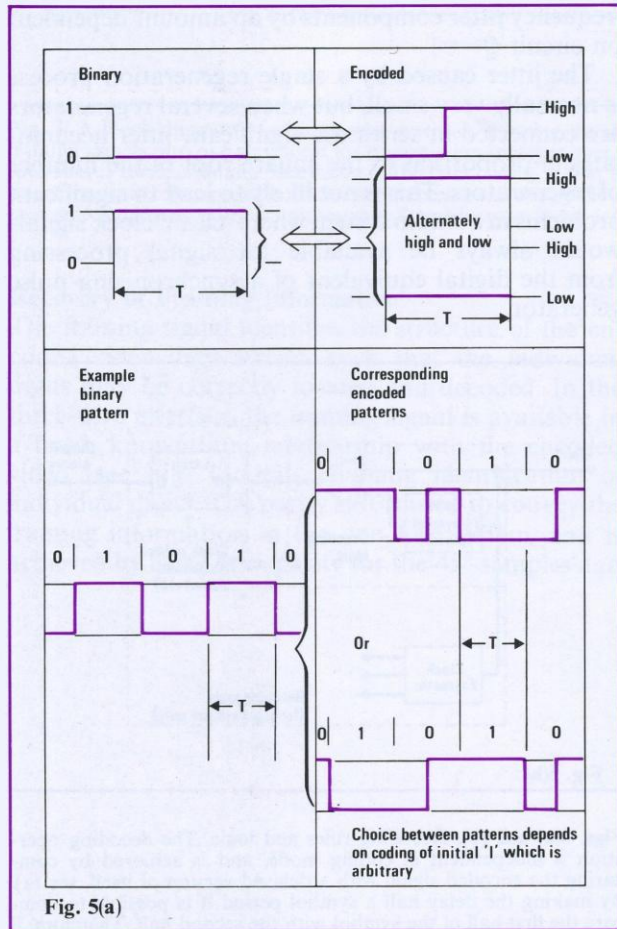
Fig. 5(a)

Figs. 5(a) and (b). Coding rules and logic. The rules for the interface coding scheme are illustrated in (a). A binary '0' is replaced by a waveform which occupies a 'low' voltage level for half a period and a 'high' voltage level for the remaining half period. A binary '1' is replaced by a waveform which remains for the whole symbol period at either the 'high' or 'low' voltage level depending upon the previous binary '1' coding. If the previous '1' was coded 'high' it will be coded 'low', and vice versa. This alternation continues throughout coding and any coded pattern can therefore exist in two possible modes. The choice between them is arbitrary, and the decoding scheme is independent of either. The logic for encoding the binary into this new form is implemented very simply, as shown in (b). The low-frequency energy of the signal spectrum is reduced by virtue of the alternating 'high' and 'low' voltages and the inherent lack of dc component in the symbol representing binary '0'.

Fig. 5(b)

recovery of data, clock and framing signals from the composite coding format.

The coding format employed is identical with that chosen for distant interfaces. The most desirable features of any coding format relevant to a single-cable interface are, the ease with which (a) the clock, and (b) the framing signals, can be recovered.

Extraction of Clock Signal

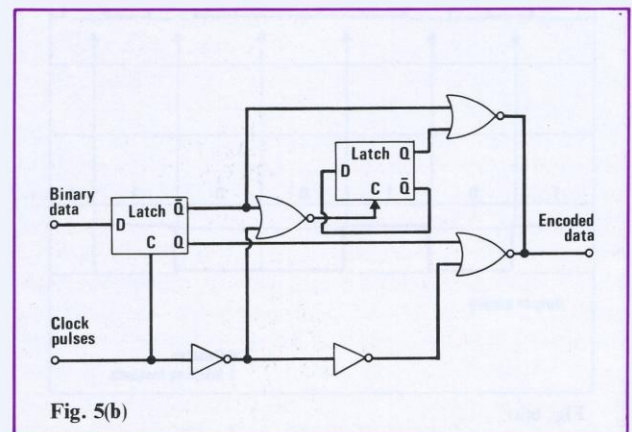It is a property of the method of encoding that, regardless of binary input patterns, negative transitions in the two-level waveform occur at least once in every two symbols. By generating an impulse every time a negative transition occurs in the waveform, and using this impulse stream to excite a tuned circuit or a phase-locked loop responsive to the binary digital repetition rate, a clock signal can be recovered. Because the excitation to the timing extraction circuit is so

regular, the amplitude and phase modulation on the extracted clock signal will be small, and will require relatively little further processing to provide a usable signal. The phase modulation, or jitter, on this recovered clock signal will depend upon the tuned-circuit Q-factor and the jitter present on the excitation driving the tuned circuit. The tuned circuit acts as a filter to the jitter spectrum and reduces relatively high-frequency jitter components by an amount dependent on circuit Q.

The jitter caused by a single regeneration process is normally very small, but when several regenerators are connected in series the significant jitter accumulates in proportion to the square root of the number of regenerators. This is not likely to lead to significant problems in a studio system where 'clean' clock signals would always be available for signal processing from the digital equivalent of a synchronising pulse generator.

Fig. 6(a)

Fig. 6(b)
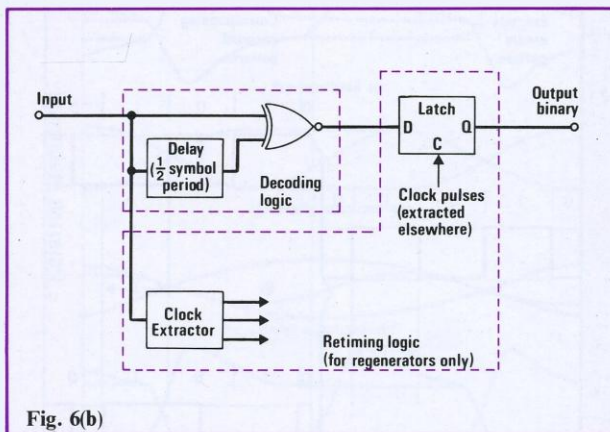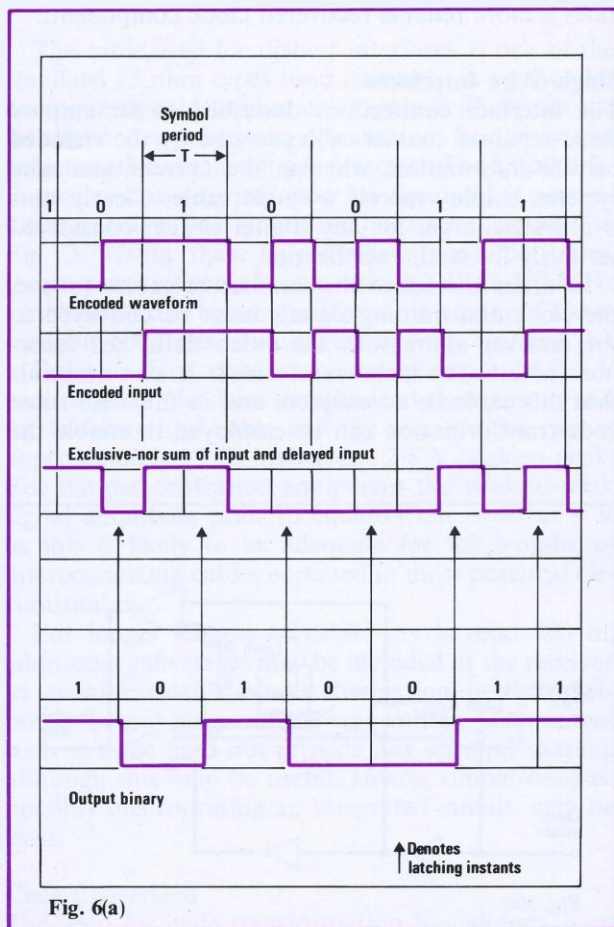
Figs. 6(a) and (b). Decoding rules and logic. The decoding operation is independent of coding mode, and is achieved by comparing the encoded signal with a delayed version of itself, see (a). By making the delay half a symbol period it is possible to compare the first half of the symbol with the second half. Therefore, if the voltage levels are different it must be decoded to give a binary '0', whilst if they are the same it must give a binary '1'. The decoding logic is thus extremely simple, as shown in (b). In general, a clock extractor is required to decode the signal, but in the 'distant' interface a separate clock signal is transmitted as well as the encoded data, hence no clock extractor is needed. Note that full regeneration would be required in the case of a single-wire interface.
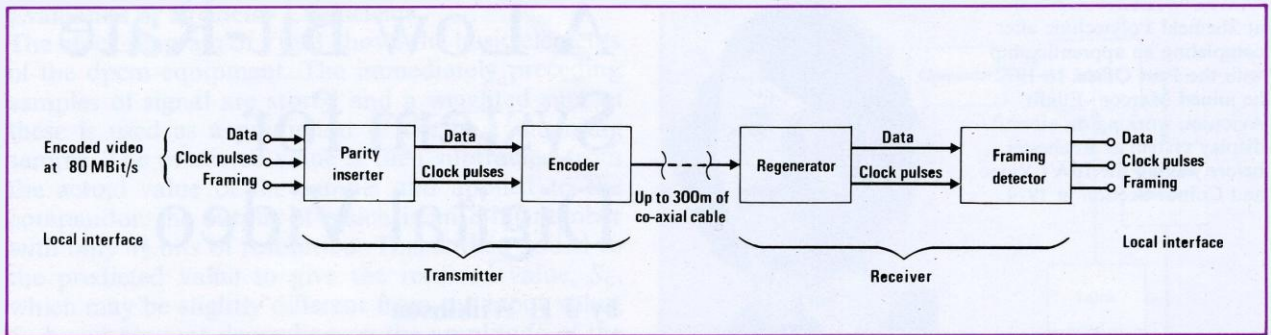
Fig. 7. A block diagram of a single-wire interface showing the processing required to enable composite coding of the data, clock and framing signals into one waveform for transmission together over a single-wire link. The parity inserter causes variation of the parity slot such that it carries framing information. The data coding is that illustrated in Figs. 5 and 6, and extraction of the clock signals from the waveform is therefore possible. Thus, at the receive end of a single-wire interface, data, clock and framing signals can be recovered from the received composite waveform although full regeneration is needed.

Recovery of Framing Information

The framing signal identifies the structure of the encoded video digit stream such that the individual digits may be correctly located and decoded. In the three-wire interface, the framing signal is available in a fixed, known-time relationship with the encoded video (see Fig. 1), thus enabling identification of individual digits. The parity slot is used to convey the framing information in the one-wire system, and is achieved by using even parity for the 45° samples and odd parity for the 225° samples. The receiver uses a correlation detector to locate the even and odd parity words. A framing signal is then generated which has the required timing relationship with the output data.

Figure 7 is a schematic diagram of the complete single-wire terminal. All the hardware is contained in one self-powered unit, and the transmit and receive sections are connected by means of an external length of coaxial cable. The equalisation may be set for any length of cable up to 300 m (1000 ft).

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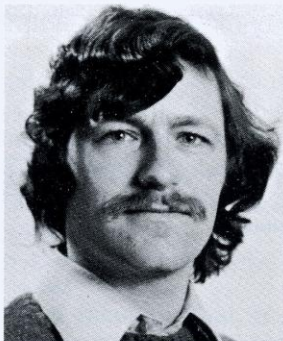
A Low Bit-Rate System for Digital Video

by J H Wilkinson

Synopsis

Recent work has demonstrated the practicality of a digital television system operating at a sub-Nyquist sampling frequency, and the principles of the $2f_{sc}$ sampling method are described elsewhere in this issue. Such a system reduces the sampling rate to a level lower than would otherwise be possible for transmission on band-limited digital systems; the effective data rate with 8-bit resolution is approximately 70 Mbit/s.

Sub-Nyquist signals are very efficient in their use of video spectrum, and have much to offer in the search for high quality at low bit rates. There is also the possibility of lowering the data rate further to 34 Mbit/s by using differential pulse code modulation (dpcm). This data rate has been adopted as a standard for international exchange of programmes within Europe. Though it is not regarded in the UK as being adequate for transmissions of the highest quality, it might be used here for news-gathering purposes. This article describes a method by which this bit rate may be achieved whilst maintaining a high quality.

The principle used is that of predicting the value of the next sample to be taken based on information already available at the receiver. The difference between the predicted value and the actual value of the sample is then compressed, i.e., the number of quantised levels is reduced prior to transmission. Sufficient information is thus available at the receiver for the original sample value to be reconstructed. Because of the compression process, improved accuracy in the prediction method will produce a higher received picture quality. Furthermore, the degradation in the received quality is mainly associated with contours in the picture, and appears as 'edge-busyness'.

The work described here is centred chiefly on the accuracy of the prediction method and the optimisation of the compression law. It explains the simple predictor constructed for the EBU demonstration and makes some suggestions for further consideration.

Introduction

Differential pulse code modulation (dpcm)[1] uses a linear predictor based on the values of the four preceding samples. It does not require the use of a PAL decoder for separating the luminance and chrominance components, but operates directly on the pcm encoded sub-Nyquist[2] data. If the coefficients of the predictor are chosen with care, it will be much more effective than predictors based on either the immediately previous or the second previous sample alone. The predicted value of a sample is then subtracted from its actual value and the resulting difference signal applied to a compressor/expander device within the transmitter. Were the equipment supplying a low bit-rate signal to a transmission link, then the data would be taken from the output of the compressor. In that case, a complementary expander device would be available in the receiver. In order to assess the impairment introduced by the dpcm equipment it is necessary only to construct a transmitter. The compressor/expander device then condenses into a single unit called the compandor. The characteristic of this is that the effective number of bits per sample is reduced from eight to four and a half. Hence, after removal of the line and field blanking intervals the data rate is reduced to below 34 Mbit/s.

This article describes the selection of a suitable predictor based on four samples, the requirements of the compandor and important peripheral elements of the equipment. It also describes the impairments introduced by the equipment as assessed by the author, and it suggests possible improvements in design.

Evaluation of Predictor Coefficients

The block diagram of Fig. 1 shows the basic elements of the dpcm equipment. The immediately preceding samples of signal are stored and a weighted sum of these is used as a prediction of the next incoming sample. The predicted value is then subtracted from the actual value of the sample and applied to the compandor, the output of which is an 8-bit number with only $4\frac{1}{2}$ bits of resolution. This is then added to the predicted value to give the received value, S_0, which may be slightly different from the input value, S_i, by an amount depending on the amplitude of the signal applied to the compandor.

Samples of a $2f_{sc}$ PAL signal consist of a luminance component added to a chrominance component which alternates in sign from sample to sample, see Fig. 2. Because of this structure, there are only certain combinations of coefficients which, when applied to the sample values, will give accurate predictions. Referring to Fig. 2, the samples numbered S_2 and S_4 have the correct phase of chrominance with respect to the sample being predicted, whereas samples S_1 and S_3 have an inverted chrominance phase. If the predictor is to be accurate in high-chrominance regions of the picture it is necessary that the sum of the coefficients allocated to samples S_1 and S_3 be approximately zero, and that the sum of those allocated to S_2 and S_4 be unity. Furthermore, in order to ensure reasonable response to transients, the coefficient of sample S_1 should be as near to unity as can be achieved. In contrast to this is the general rule that the sum of the moduli of the coefficients should be small in order to reduce overall noise. The problem of predictor design is to find the best compromise between these conflicting requirements. Predictors using only three previous samples are possible, but are unsatisfactory in performance, while sample positions of fifth and higher orders are too far removed from the position of the sample being predicted to be of any value. This reasoning led to the choice of predictors based on four samples.

There are two simple methods of determining the performance of different types of predictor. One is to calculate the frequency response of the signal applied to the compressor, the other is to derive an approximate measure of the predictor transient response using a $2T$ pulse and a positive-going edge as inputs. Various predictor coefficients have been considered and the best results for each test are displayed in Figs. 3 and 4. The two sets of coefficients which give best theoretical results are (i) $\frac{1}{2}, \frac{1}{2}, -\frac{1}{2}, \frac{1}{2}$ and (ii) $\frac{3}{4}, \frac{1}{2}, -\frac{3}{4}, \frac{1}{2}$. Because of uncertainty as to which of these
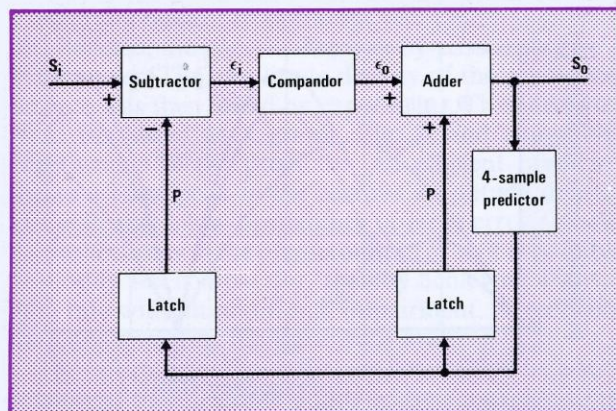
Fig. 1. The basic elements of a differential pulse code modulation (dpcm) encoder. The predictor makes an estimate, p, of the value of the next incoming sample from the four previous samples in time. The accuracy of this estimate determines the value of the prediction error, ϵ, applied to the compandor; the better the estimate, the smaller the value of ϵ_i. The errors introduced by the compandor, i.e., $\epsilon_0 - \epsilon_i$ are increased as the magnitude of ϵ_i increases. This results in distortions being added to the output signal S_0. The level of distortion is a function of the predictor algorithm and the amount of compression introduced by the compandor. A compandor is the consolidation of two processes—compression and expansion.
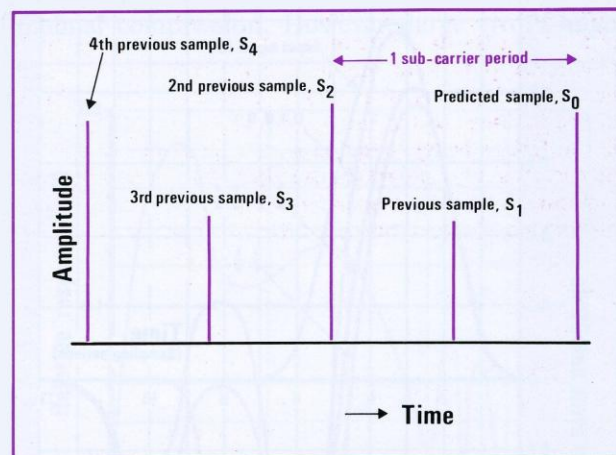
Fig. 2. The magnitudes of samples of typical $2f_{sc}$ data are shown here. The values of the predictor coefficients are selected to overcome the variation of signal amplitude which occurs between consecutive samples due to the specified relationship between sub-carrier frequency and sampling frequency.

would give the better result, the predictor was designed to enable selection of coefficient values, whereby the final decision could be made subjectively. By using standard arithmetic logic units in five different operational modes it is possible to select coefficient values of 0, $\frac{1}{4}$, $\frac{1}{2}$, $\frac{3}{4}$ or 1 for each sample position. Another approach which can yield useful
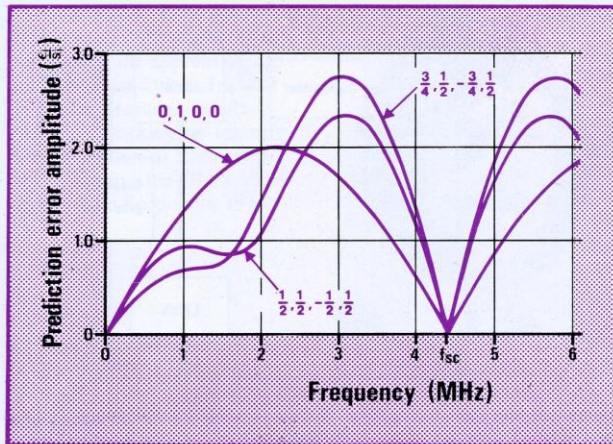
Fig. 3. The frequency response of three types of dpcm encoder for which the predictor coefficients are shown against their respective curves. Each response has been calculated from the Fourier transform of the predictor values. The shape of each curve determines the visual quality of the output picture. Ideally, this should be inversely matched to the spectrum of the input signal for the minimum mean square value of ϵ_i (see Fig. 5).
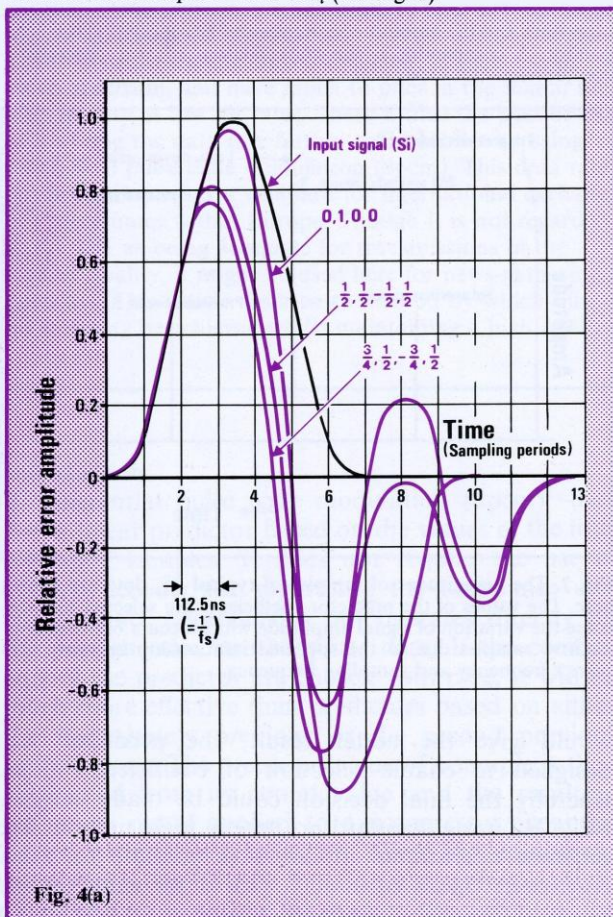
information about predictor performance is to examine the product of the frequency responses shown in Fig. 3 multiplied by the spectrum of the signal input to the system. If the resulting graph is reasonably flat it can be assumed that the predictor is efficient. Such a performance is indicated by the substantially flat response shown in Fig. 5. It should be noted that it is possible to flatten the response completely, but to do this requires additional samples which, it is believed, would probably not appreciably improve the overall quality.

A mathematical technique for determining the values of the coefficients is given in the Appendix. The values are obtained by exploring the correlation between the sample being predicted and the four previous samples. This gives a set of coefficients which are dependent on the average ratio between the chrominance and luminance signal components. For typical values of chrominance/luminance ratio, this method
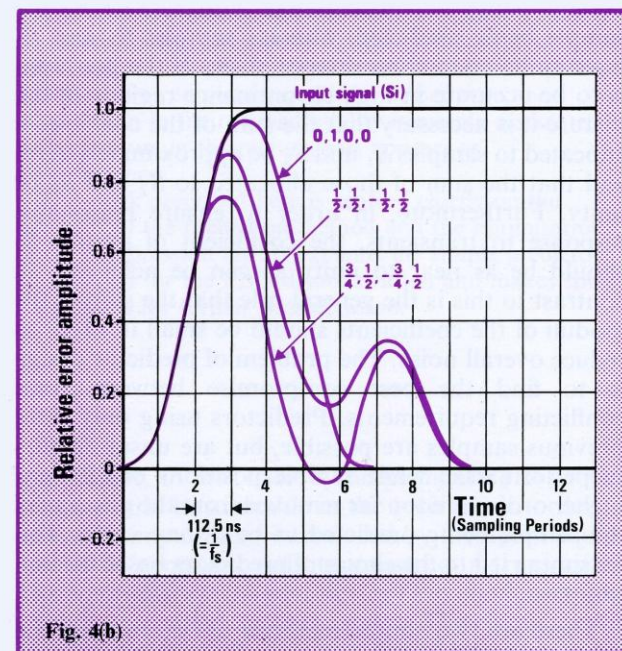
Fig. 4(a)

Fig. 4(b)

Figs. 4(a) and (b). The calculated transient response of three types of dpcm encoder. Fig. 4(a) shows the input signal, S_i, as a simulated $2T$ pulse. All dpcm encoders produce ringing, and this will be reflected in the output picture as 'edge-busyness'. The multiple-sample predictors are selected to minimise the ringing effect, as compared with the second (only) previous sample predictor (0, 1, 0, 0). Fig. 4(b) shows the responses of the predictors to the positive edge of a simulated bar.

produces substantially the same coefficient values as the methods described above.

System Block Diagram

The block diagram of the equipment is shown in Fig. 6. There are four features of the system which require further explanation since they are additions to the basic process illustrated in Fig. 1.

i By means of a 3-way switch the output of the system can be selected between (a) pcm, (b) dpcm and (c) the difference between pcm and dpcm. The switch is normally moved between positions (a) and (b) to gain a visual impression of the impairment introduced by the dpcm process, and switch position (c) is used for making an objective measurement of that impairment.

ii The characteristic of the companding device can be changed between two alternatives by means of a second switch. This permits a direct comparison to be made between two different compandor characteristics and enables that which is optimum to be selected experimentally.

iii A third switch enables selection between two adjustable sets of predictor coefficients, and is very useful for the making of comparison tests. Each sample coefficient of both sets is independently variable and can be given any of the values $0, \frac{1}{4}, \frac{1}{2}, \frac{3}{4}$ and 1. The third coefficient is always negative because, as stated earlier, the sum of the first and third coefficients must be zero, and the first coefficient is always positive. The second and fourth coefficients are both always positive; no practical predictor is likely to have negative coefficients in either of these positions.

iv The input signal is compressed to reduce the number of possible input levels from 256 to 208. This prevents the possibility of overloading which could otherwise occur when using modulo 256 arithmetic in dpcm equipment. Though this compression process is reversed at the output, it leaves a loss of resolution in the extremes of the video range, see Fig. 7. It has been arranged that the compression affects only those signals in the region of peak white luminance level or below the negative peaks of the colour burst. In these compressed regions, the resolution is reduced from 8 to 7 bits, but all normal picture information is left unimpaired.

Measurements of Picture Statistics using the Selected Predictor Coefficients

A series of subjective tests established that the set of

coefficients nearest to optimum was $\frac{1}{2}, \frac{1}{2}, -\frac{1}{2}, \frac{1}{2}$. With this decided, the compandor was by-passed to give a 1:1 transfer. The amplitude density of the prediction error, ϵ, was then found by experiment. This measurement is plotted as a graph, Fig. 8, and shows the effectiveness of the predictor. An efficient predictor shows a large peak in the relative amplitude density at $\epsilon = 0$, and a low probability of large error signals. As the accuracy of the predictor improves, it is possible to modify the compandor, thereby achieving a lower data rate without additional impairment, or reducing the impairment at the same data rate.

The optimum compandor was determined by a succession of modifications followed by subjective tests.

Factors Influencing the Shape of the Compandor Characteristic

The compandor device consists of two programmable read-only memories (prom) each with eight address inputs and four outputs. For reasons of speed, the compression and expansion were combined as a single operation within the compandor. The nature of the characteristic is evident in Fig. 9 (see p. 56). Small prediction errors pass through the system with minimal compression. However, large errors might
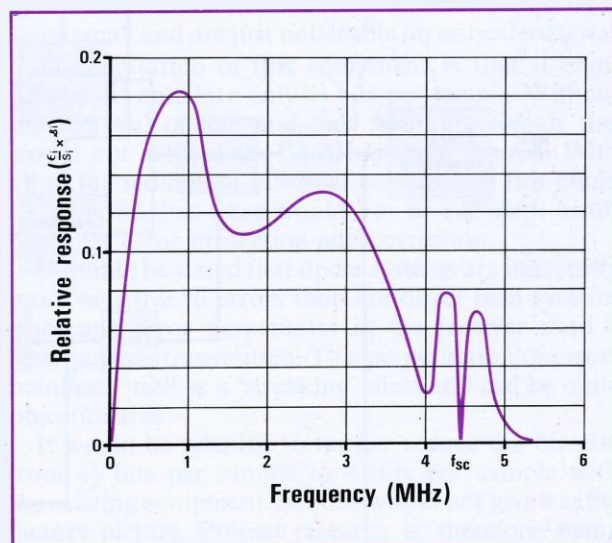
Fig. 5. This graph shows the effect of multiplying the frequency response $(\frac{1}{2}, \frac{1}{2}, -\frac{1}{2}, \frac{1}{2})$ of Fig. 3 by the envelope of the spectrum of the incoming signal, σ_i. The spectrum was derived from the mean of five slides prepared by the EBU for subjective assessment of digital television systems. According to classical dpcm analysis an optimum encoder should exhibit a flat response, but in reality this is impracticable. Further, the analysis does not consider the effects of gamma or subjective assessment. Using a degree of interpretation, however, this response can be useful in indicating some of the effects that the encoder is likely to have.

suffer considerable compression. The figure shows the 22 possible states which are represented by $4\frac{1}{2}$ bits at *the compressor output, and that the compression is achieved by coding a series of input levels to a single* output level. In general, low values of prediction error (which occur most frequently), are coded with the highest resolution. The limit to the compression that can be applied to the range of input levels 80–155 is determined by the fact that the largest compression error generated must be no greater than 24 to prevent the possibility of overloading in the receiver.

The compandor characteristic represents a compromise between the requirements of 'edge-busyness' and background noise. If the requirement is that of reducing the 'edge-busyness', high-level resolution can be improved at the expense of low-level resolution. However, it has been found that in general the best policy is to ensure a quiet background at the expense of 'edge-busyness'. This is because typical pictures have very few transients; and low-level information, e.g., in facial contours and in hair, should be preserved with highest quality.

The shape of the compressor characteristic shown in Fig. 9 represents a good compromise in the interests of maintaining good overall picture quality.

Practical Aspects of the DPCM System

The dpcm process must be implemented using parallel data format since the compandor devices are available only in this form. This, together with the high speed required of the dpcm circuit dictates that emitter-coupled logic, ecl, must be used. The only exceptions are the compandor devices which use transistor-transistor logic, ttl. Comparable ecl devices have since become available and these also would be suitable. The time available for the necessary calculations within the dpcm loop is 112 ns. In that time the 8-bit parallel data must pass through four
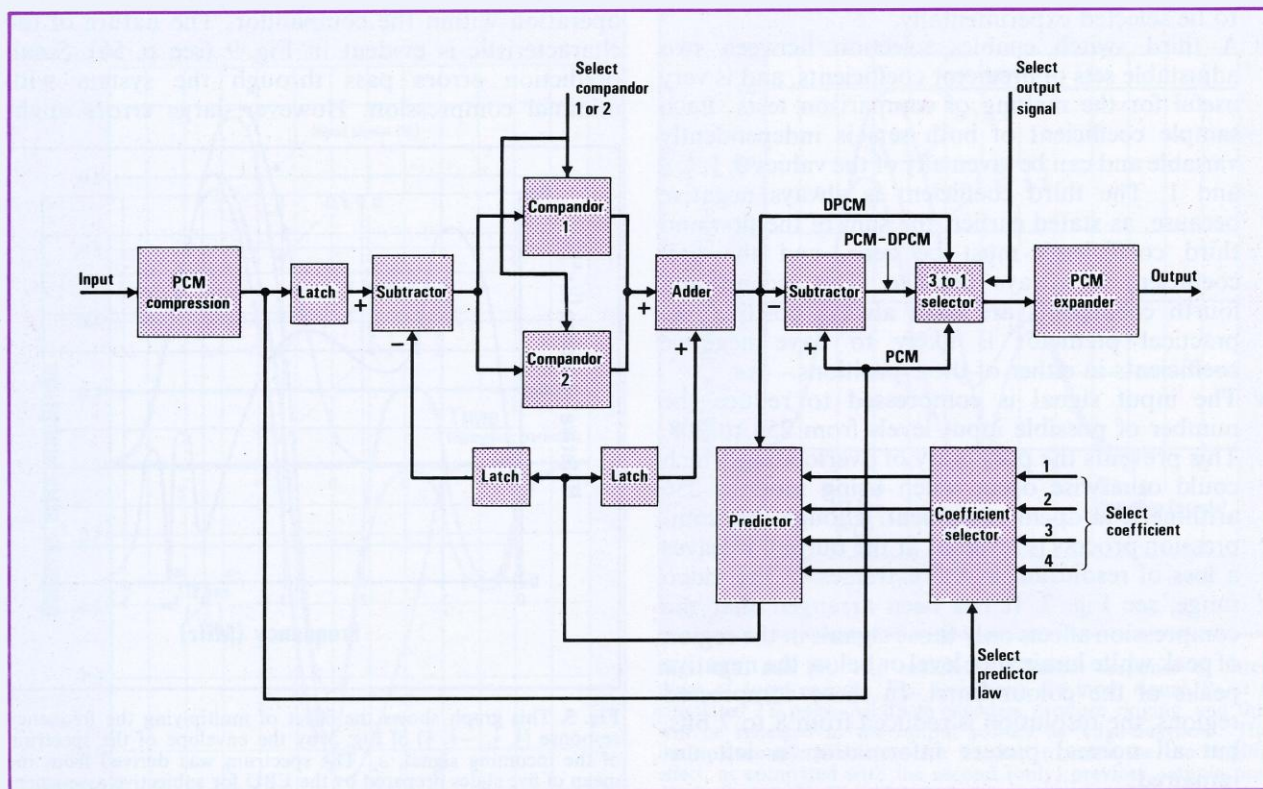
Fig. 6. The block diagram indicates the practical realisation of the scheme outlined in Fig. 1. There are four additional facilities; (i) input signal compression to eliminate the possibility of overload, (ii) selection between two compandors to assist in the optimisation of the companding characteristic, (iii) selection between different predictors to confirm the theoretical work (each coefficient value can be adjusted from 0 to 1 in steps of 0·25), and (iv) output signal selection between S_i, S_0 and $S_i - S_0$. Switching between S_i and S_0 gives a quick visual display of the impairments introduced, and the last position, $S_i - S_0$, allows an objective measurement of the impairments.

Fig. 7. Using a scaling factor of 5 mV/quanta for video, and taking 100% colour bars as the maximum signal to be handled, the quantised levels corresponding to the different parts of the waveform are as shown 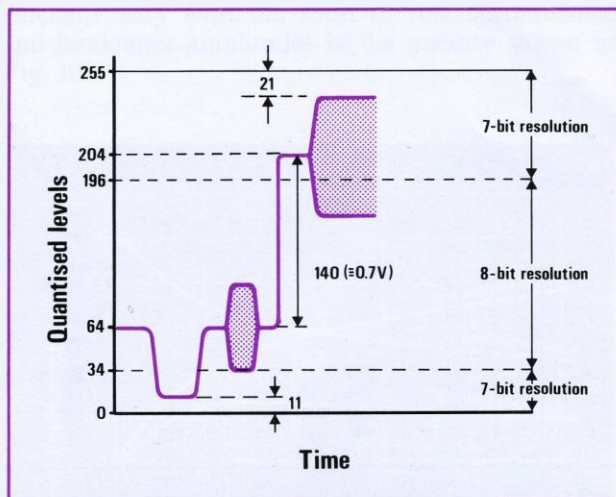in the diagram. Prevention of overload of the dpcm encoder requires compression of the input signal range to 208 levels (centred on level 128). This has been achieved by changing the resolution of the extremes of the signal to 7 bits per sample and shifting the remaining numbers to achieve symmetry about level 128. At the output the process is reversed.
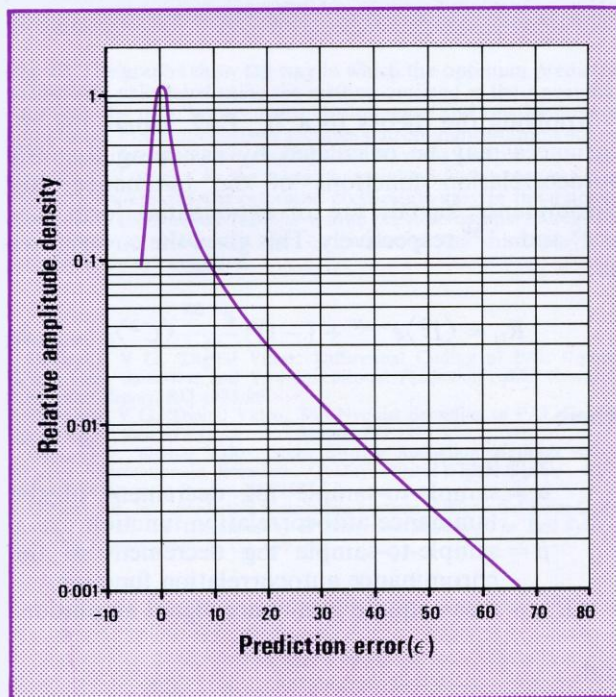

Fig. 8. This graph shows the relative amplitude density of the values of ϵ (with no companding, $\epsilon = \epsilon_i = \epsilon_0$). The object of the dpcm encoder is to obtain as narrow a distribution as possible. This information is then interpreted to derive the companding characteristic.

arithmetic operations, the compandor devices, two latching operations, the ecl-to-ttl and the ttl-to-ecl conversions. Consequently, there is very little time available for each individual operation, and great care has been taken in the layout of each board to ensure high speed. The boards used are of the earth plane type with the power supplied by a matrix of pylons. The supply to every integrated circuit is decoupled to the earth plane to ensure a low value of supply voltage transients. The layout was carefully arranged to render the signal paths as short as possible.

Conclusions

This approach to dpcm results in a very simple set of predictor coefficients. With the exception of certain test material the pictures show little impairment. The impairments appear as noise on high level transients and in areas where there is a high-amplitude component in the region of 2 MHz. This becomes plainly evident on the gratings of test card F. The only other significant impairment is a slight increase in overall picture noise especially on pictures having large areas of low-level detail. A particular example of this is the EBU test slide, 'The Tree'. It must, however, be stressed that these impairments are quite small and are just noticeable on critical material.

The limitation of this equipment is that it compresses the data into only $4\frac{1}{2}$ bits per sample. Without the removal of line and field blanking signals this would not achieve the 34 Mbit/s rate desired. With all of the redundant information removed this could be achieved, but there would not be sufficient headroom for error protection and correction.

It should be stated that dpcm systems are inherently more sensitive to errors than are direct pcm systems since any error perpetuates in the receiver until a reset pulse is transmitted. This perpetuation of errors manifests itself as a 'streaking' effect and can be quite objectionable.

It would be possible to further reduce the bit rate from $4\frac{1}{2}$ bits per sample to 4 bits per sample with the existing equipment, but this would not give a satisfactory picture. Present research is, therefore, being undertaken in an attempt to improve the dpcm process in two ways. First, by designing an adaptive predictor which avoids using samples associated with a boundary. This should reduce the effect of 'edge-busyness'. Secondly, by developing a more efficient coding strategy. Variable-length codes exist which reduce the data rate without causing any picture impairment.[3] Unfortunately, such codes are very difficult to use,
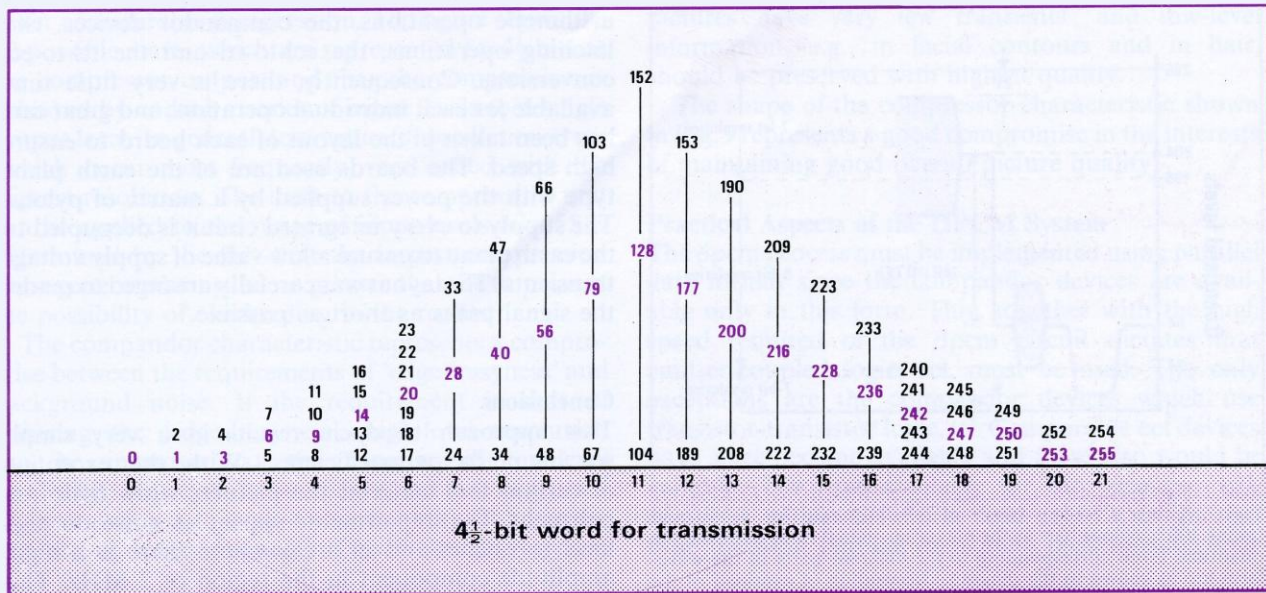
Fig. 9. The companding characteristic. Any digital number from 0 to 255 applied to the input of the programmable read-only memory, will appear at the output as one of only 22 numbers. Referring to the diagram, when any number in a vertical column is applied to the input of the compandor, the output will be the number in that column shown coloured, i.e., 16 at the input will appear at the output as 14. Each of these 22 numbers is then encoded to form a 4½-bit word ($2^{4 \cdot 5} = 22 \cdot 63$), as shown in the tinted area, which is available for transmission. In practice the transmitted data would be available in groups of 9 bits, each group containing two 4½-bit words.

but it might be possible to combine some of the advantages of both variable-length and non-linear codes.

In view of the further development work currently taking place in this sphere within the IBA, it is not envisaged that formal viewing tests will be made using the equipment described here, but it is available as a reference for future work.

Appendix

With the object of optimising the coefficient values of the predictor with respect to the rms prediction error, the method employed makes use of the linear prediction theory originated by O'Neal.[4] This shows that for a prediction of the form:

$$p = \sum_{i=1}^{m} a_i x_i \quad \ldots \quad \ldots \quad \text{(A.1)}$$

the minimum rms prediction error occurs when the coefficients (a_i) are chosen to satisfy a set of simultaneous equations in the covariances $R_{ij} = \overline{x_i y_j}$:

$$R_{0j} = \sum_{i=1}^{m} a_i R_{ij} \quad (j = 1 \cdots m) \quad \ldots \quad \text{(A.2)}$$

Applying the theory to a $2f_{sc}$ PAL signal, the covariances may be calculated by assuming that the autocorrelation functions of the luminance and chrominance signals are of exponential form, i.e., $e^{-\alpha N}$ and $e^{-\beta N}$ respectively. This gives the covariances as:

$$R_{ij} = \langle L^2 \rangle e^{-\alpha N} + (-1)^N \frac{e^{-\beta N}}{2} \langle C^2 \rangle, \quad \text{(A.3)}$$

where

$N = |i - j|$
α = sample-to-sample log decrement of the luminance autocorrelation function
β = sample-to-sample log decrement of the chrominance autocorrelation function
$\langle L^2 \rangle$ = mean square luminance signal amplitude

and

$\langle C^2 \rangle$ = mean square chrominance signal amplitude.

The Equations (A.2) can be solved using the covariances of Equation (A.3). Optimum predictor co-

efficients vary with the ratio of rms chrominance and luminance amplitudes in the manner shown in Fig. 10.
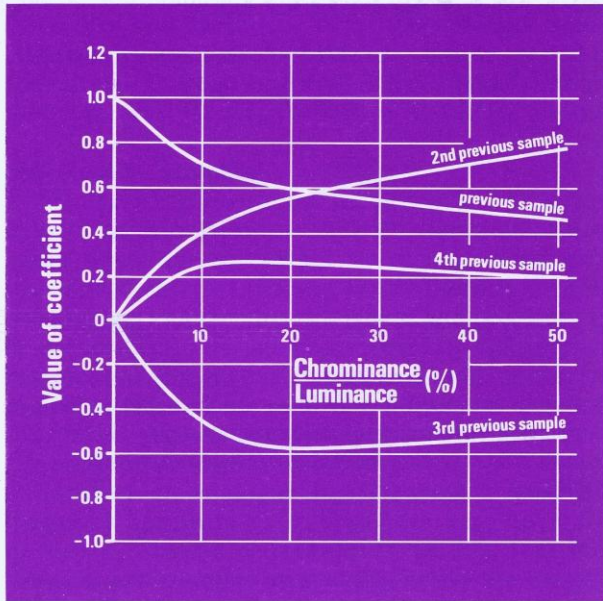
Fig. 10. The graphs show the way in which the optimum predictor coefficients, calculated using the method outlined in the Appendix, vary with the ratio between rms chrominance and rms luminance amplitudes. For monochrome pictures (chrominance/luminance = 0) the optimum predictor employs the previous sample to the exclusion of all others. As the mean chrominance/luminance ratio in the picture increases, the optimum coefficients vary in the manner shown, with the coefficient of the second previous sample increasing in importance. Note that the coefficient of the third previous sample is always negative.

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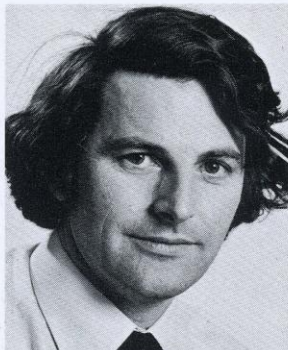
Analogue-Digital Conversion

by I R Lever

Synopsis

There are two fundamental techniques available for the generation of digital signals sampled at sub-Nyquist rate. The first is the obvious approach of operating the sampling device at the sub-Nyquist rate and then minimising the consequent picture degradation produced by aliasing using analogue filtering techniques at the analogue/digital interface. Such filters, although apparently simple to design, are potentially liable to drift causing an increased rather than a reduced amount of degradation.

The alternative is to sample the incoming video signal above the Nyquist rate, and then to use digital techniques for filtering and changing to the required sub-Nyquist format, thus ensuring definable precision and high stability.

Introduction

By processing television pictures in the digital format it is possible for signals to be manipulated with predictable and repeatable precision. The price that has to be paid for this is a considerable increase both in circuit operating speed and in signal bandwidth.

Until mid-1976 virtually all the digital equipment designed by the IBA for television purposes sampled the encoded PAL signal at three times sub-carrier frequency which, using an 8-bit word per sample, yields a bulk data rate of 106·4 Mbit/s and requires rather more than 50 MHz of bandwidth compared with 5·5 MHz for the original analogue signal.

As the costs involved in signal processing are generally directly related to the speed of operation and the bandwidth required, one of the objectives in constructing the experimental sub-Nyquist studio system was to evaluate the relationship between picture quality and reduced data rate, and hence the potential cost of such a system.

Digitisation

In order to convey information in digital form, two main parameters must be determined. These are the number of 'bits' used for describing the magnitude of each sample, and the rate of sampling.[1] A system of 8-bit linear (256 level) quantisation has become almost standard practice in digital television operations and was adopted in this equipment. As already mentioned, the sampling rate used in earlier work had been $3f_{sc}$. This rate had been chosen for the following reasons:

1. The rate had to comply with the criterion, first noted by Nyquist, that for sampling an analogue signal such that it may be reconstructed from the derived samples without error, the samples must be generated at a rate greater than twice the highest frequency contained by the signal.

2. The sampling should be coherent and simply related to the colour sub-carrier. The physical process of sampling is never perfect; some non-linearity will always be present giving rise to intermodulation between the sampling and the signal appearing as a patterning (similar to moiré) in coloured areas of the viewed picture. If the sampling is coherent the patterning remains stationary with respect to the sub-carrier and is therefore considerably less noticeable.

3. The sampling rate had to be within the capabilities of the analogue-to-digital converters (adc)

that were then available, placing an upper limit at approximately 15 MHz. At that time an adc capable of this level of performance cost approximately £2200.

Hence, the sampling rate had to be greater than $2 \times 5\cdot5 = 11\cdot0$ MHz (i.e., twice the analogue signal bandwidth), less than 15 MHz and simply related to the sub-carrier frequency, 4·43 MHz. The only choice, therefore, was $3f_{sc}$, i.e., $3 \times 4\cdot43 = 13\cdot3$ MHz approx.

Sub-Nyquist Sampling

If the sampling operation be performed at a rate below that stipulated by Nyquist, the original analogue signal can no longer be reconstructed from the derived samples without error. Distortion products, known as aliasing components, will be generated in association with all high-frequency information beyond the Nyquist criterion, and will appear as moiré patterns in areas of fine horizontal detail in the viewed picture. The mechanism by which these are produced is explained below.

Consider an analogue signal having a baseband frequency spectrum extending from zero to a frequency, f_b, where it ends abruptly, see Fig. 1(a). This signal is now sampled at a rate, f_s, above the Nyquist limit, yielding a train of very narrow pulses of magnitude corresponding to the instantaneous amplitude of the analogue waveform at each sampling point, see Fig. 1(b). This results in the spectrum shown in Fig. 1(c). It will be seen that in addition to the original spectrum additional new components have been generated about the sampling rate, f_s. Similar components, not shown in the diagram, are also generated about each integral multiple of f_s extending towards infinity.

The original baseband signal may be recovered without distortion, other than a considerable loss in amplitude, by passing the pulses through a low-pass

Fig. 1. The stages in sampling and recovering an analogue waveform. The idealised frequency spectrum of the baseband analogue waveform is represented at (a). This signal is then sampled at a rate, f_s, to produce a series of narrow pulses of information, (b). The result of this is that, in addition to the original spectrum, a translation of the baseband spectrum appears on either side of f_s as shown at (c). The original waveform is recovered by means of a low-pass filter as in (d), but (e) shows that at the Nyquist limit, where $f_s = 2f_b$, the idealised spectra are contiguous. In practice, however, due to the finite rate of filter roll-off, they overlap as at (f), and the baseband signal can then no longer be simply filtered out without including the aliasing components, shown shaded. In the case of a PAL television signal sampled at $2f_{sc}$, shown at (g), the aliased area is considerable, and the baseband signal may be recovered only by means of comb filtering techniques.
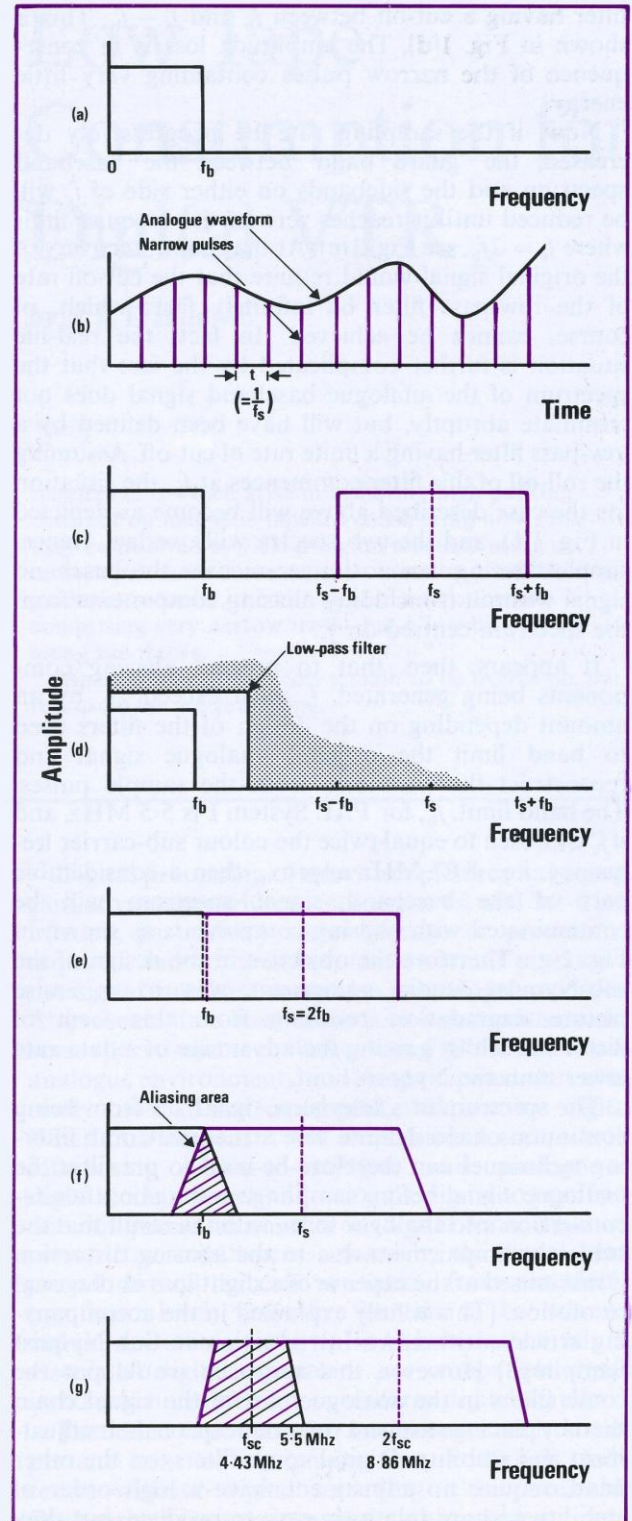
filter having a cut-off between f_b and $f_s - f_b$. This is shown in Fig. 1(d). The amplitude loss is in consequence of the narrow pulses containing very little energy.

Now, if the sampling rate be progressively decreased, the 'guard band' between the baseband spectrum and the sidebands on either side of f_s will be reduced until it reaches zero at the Nyquist limit where $f_s = 2f_b$, see Fig. 1(e). At that point, recovery of the original signal would require that the cut-off rate of the low-pass filter be infinitely fast; which, of course, cannot be achieved. In fact, the real-life situation is further complicated by the fact that the spectrum of the analogue baseband signal does not terminate abruptly, but will have been defined by a low-pass filter having a finite rate of cut-off. Assuming the roll-off of this filter commences at f_b, the situation for the case described above will become as depicted in Fig. 1(f), and the two spectra will overlap. Hence, simple filtering can no longer recover the baseband signal without it including aliasing components from the spectrum centred on f_s.

It appears, then, that to prevent aliasing components being generated, f_s must exceed $2f_b$ by an amount depending on the design of the filters used to band limit the original analogue signal and reconstruct the waveform from the sample pulses. The band limit, f_b, for PAL System I is 5·5 MHz, and if f_s is chosen to equal twice the colour sub-carrier frequency, i.e., 8·87 MHz approx., then a considerable part of the baseband signal spectrum will be contaminated with aliasing components as shown in Fig. 1(g). Therefore the objective in the design of the sub-Nyquist studio equipment was to minimise picture degradation resulting from this form of distortion whilst gaining the advantage of a data rate lower than the Nyquist limit.

The spectrum of a television signal, far from being continuous, has a definite 'line' structure. Comb filtering techniques can therefore be used to pre-filter the analogue signal before sampling, and again after re-conversion into analogue form, with the result that the subjective impairment due to the aliasing distortion is minimised at the expense of a slight loss in diagonal resolution. (This is fully explained in the accompanying article entitled 'An Introduction to Sub-Nyquist Sampling'.) However, this approach would put the comb filters in the analogue part of the signal chain thereby placing stringent requirements on their adjustment and stability. Digital comb filters, on the other hand, require no adjustment, have a high order of stability and are relatively easy to produce, but they

can be used only after the analogue signal has been sampled and quantised. To avoid aliasing problems, the initial sampling rate must occur above the Nyquist limit, and, as the resultant signal is required at $2f_{sc}$ rate, the most convenient choice for this is $4f_{sc}$. Digital filtering and conversion to $2f_{sc}$ can then take place in a single operation as explained in the companion article on page 21 entitled 'Digital Sub-Nyquist Filters'. For converting back into the analogue format, the above procedure is reversed. The conversion of the data from $2f_{sc}$ to $4f_{sc}$ is combined with the filtering process, and this is followed by digital-to-analogue conversion at the $4f_{sc}$ rate.

The use of this technique required that both the analogue-to-digital and the digital-to-analogue converters operated at $4f_{sc}$, i.e., approximately 18 MHz. This would have presented quite a stumbling block, particularly in the analogue-to-digital case, as the maximum conversion speed of the devices then available was 15 MHz. Fortunately, however, the manufacturer of the only one readily available was able to upgrade this product for operation at $4f_{sc}$ rate, though the approximate cost of each was £2400. This situation prevailed until mid-1977 when the market opened up with a number of competitive devices from several sources. At the present time (mid-1978) these include at least one source of single-chip monolithic devices costing approximately £350 each.

The Future

The sub-Nyquist, $2f_{sc}$, method of handling digital PAL composite signals requires analogue-to-digital and digital-to-analogue converters that, for the filtering and conversion system used in the studio equipment, are capable of operating at speeds higher than those used in earlier $3f_{sc}$ systems. This poses no technological limitation; and, with the steady downward trend of costs, no real penalty is incurred by the higher speed requirement. One manufacturer has even predicted that, by the early 1980s, the cost of a suitable analogue-to-digital converter might be as low as £20.

Reference
1. Lever, I R, and Connolly, W P, 'Analogue Processing and Operational Controls', *IBA Technical Review 8*, September 1976, 16–30.

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Low Tape Consumption for Digital VTR

by J L E Baldwin

Synopsis

It is anticipated that analogue equipment now used in studio centres and for transmission throughout networks will be progressively replaced by the digital equivalent, thus offering the advantages of precision, consistency of performance and simplicity of operation. However, this change could not have been envisaged until digital video recording had been proved feasible.

In this article a comparison is made between the essential characteristics of analogue and digital recorders with particular emphasis placed on the problems and requirements which arise in achieving high packing densities on magnetic tape. In considering how these may be resolved to yield a digital machine using no more tape than a current 1-inch broadcast, helical-scan, analogue recorder, theory calls for a tape format comprising very narrow tracks but a low bit density along the tracks.

Many practical problems remain to be overcome, but results so far are promising.

Introduction

For more than twenty years the quadruplex video tape recorder has reigned supreme, and has become a world standard. Broadcast-quality, segmented, helical recorders using 1-inch tape have been in use for several years, and this standard is now recognised as the SMPTE Type B Format. Non-segmented, broadcast-quality helical recorders have been in use for about a year and originally were of two different types. Recently the differences between these two machines have been resolved resulting in a new standard, the SMPTE Type C Format. Machines complying with this standard should soon become available. All of these recorders use frequency modulation and, because the carrier frequency is low, moiré patterning is inherent in the system. Normally, this is visible only as a beat pattern in those parts of the picture for which the signal contains high-frequency components of large amplitude. It is particularly a problem with highly saturated 625-line PAL colour pictures due to the presence of a strong sub-carrier signal used for conveying the chrominance information.

Digital processing of television pictures has significant advantages within studio centres, and for transmission, particularly where storage or precision is required. Indeed, it is becoming normal practice to use digital time-base correction in analogue recorders. At present, time-base correctors, synchronisers and some special-effects equipment are digital islands in an analogue environment, but it is anticipated that eventually most of the equipment will become digital. When this happens it will be possible to remove the intervening digital-to-analogue and analogue-to-digital codecs thereby allowing the signal to pass from one equipment to another in digital form. This will prevent the impairment, caused by the signal being passed through successive codecs, from reaching a significant level: however, for this to be achieved within a studio centre, it is essential that all the associated digital equipment use compatible digital standards or, preferably, a uniform digital standard.

The video tape recorder has become such an important item of equipment that a digital studio centre without digital recording is inconceivable. Obviously there is no insurmountable problem in pro-

61

ducing a digital recorder, but if possible the tape consumption should be no greater than that of a current 1-inch broadcast helical-scan analogue recorder.

Comparison between Analogue and Digital Recording

Figure 1 shows a very simple block diagram of an analogue recorder with digital time-base correction. Due chiefly to mechanical imperfections the signal recovered from tape cannot be used directly because its timing is not sufficiently precise. In the case of helical recorders this timing inaccuracy could be of the order of a few microseconds and must be reduced to about one nanosecond. Theoretically, this is achieved by sampling the signal in an analogue-to-digital converter (adc) by clock pulses derived from the colour burst recovered from tape, writing the resultant words into a buffer store from which they are later read at precise time intervals, and then converting back to an analogue signal by means of a digital-to-analogue converter (dac).

Unfortunately, the colour burst recovered from tape is inevitably contaminated with noise. In itself, this would not be a problem if the timing errors changed at a sufficiently slow rate to allow the use of smoothing for substantially removing the noise; but, some timing errors arise abruptly, and this very much restricts both the type, and the amount, of smoothing that can be used. Consequently, the sampling in the adc is perturbed from its correct time-relationship

Fig. 1. The figure shows an analogue vtr which employs a digital time-base corrector for removing timing errors introduced by the tape transport. The analogue signal recovered from tape is sampled and read into a buffer store. The sampling clock pulses are derived from the colour burst in the recovered signal itself, and are therefore subject to the same timing errors as the video information. The samples are subsequently read from the buffer store using reference clock pulses, and this effectively removes most of the timing errors introduced by the recording process. The system is imperfect because the noise contained in the recovered colour burst cannot be completely removed by filtering.

with the video signal being sampled which results in the timing correction being imperfect. The buffer store itself is inherently perfect; the words read from it are identical with those written into it. It is therefore said to be perfectly 'transparent'.

The block diagram of a digital recorder is shown in Fig. 2. As before, the sampling in the adc is derived from the colour burst; but, in this case, although the signal may at times be noisy, it will not have abrupt timing perturbations. These requirements are essentially those of a sub-carrier locked oscillator in, for example, a colour monitor.

After some form of code conversion the digital words from the adc are recorded on to the tape. On replay the words read directly from tape pass through a complementary code converter, and from this the original words are recovered and written into the buffer store. However, prior to recording, some special start words are regularly added. These are used for defining the start of address sequences before they are written into the buffer store. The code recorded on tape must serve to ensure that an adequate clock component can be recovered. The information is read from the buffer store as in the case of the analogue recorder, but with the digital equipment the numbers representing the output are identical with those previously recorded (except perhaps when dropouts occur), and hence the entire process of recording and time base correction is transparent.

By extending the digital system upstream and downstream the transparent area may be increased as far as required. In addition to the inherently perfect timebase correction, a digital recorder in a digital environment does not produce any impairment other than drop-outs, and these can be either corrected or concealed more effectively than is normal practice with analogue recorders.

The penalty that has to be paid to secure these advantages is that the bandwidth required is about four times that needed for analogue recording. On the other hand, the signal/noise ratio may be considerably relaxed; a 20 dB ratio of peak-to-peak signal/rms noise would give an error rate of about 1 in 10^7. This less stringent noise requirement might therefore enable reduction of tape consumption per unit bandwidth, such that the hourly consumption of tape would be no greater than that of analogue recorders.

High-Density Recording

The playing time of a digitally recorded tape will be directly proportional to the number of bits per unit area, assuming that the bit-error rate remains con-
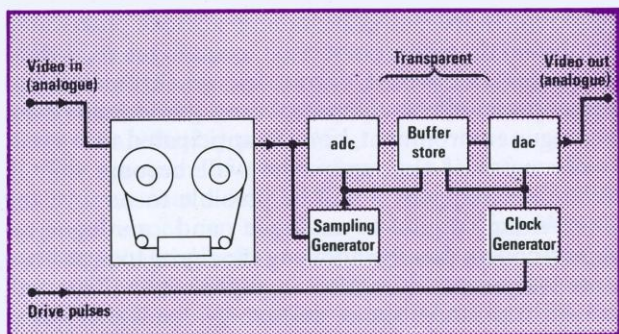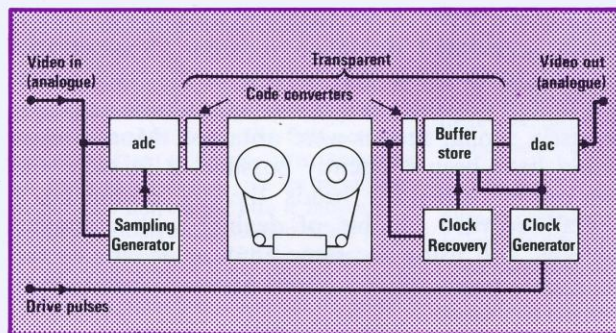
Fig. 2. In the case of the digital vtr shown here, the analogue signal is sampled at the input using clock pulses again derived from the colour burst. In this case the timing errors are very small because the burst has not passed through the recording process. Coding systems are used which allow continuous recovery of clock pulses on replay. Therefore, the recorder itself is entirely free of timing errors.

stant. The number of bits per unit area is the product of the number of bits per unit distance along a track and the number of tracks per unit distance measured across the tracks. With a reduced track width, since the number of particles influencing the pick-up head is proportionately less, the signal level and the tape noise *power* are correspondingly reduced. Assuming all the noise originates from tape, it follows that the signal/noise ratio is proportional to the square root of the track width.

Consider now a replay head. Ignoring edge effects, the inductance of a head is substantially proportional to the width of track with which it is used. For a constant value of inductance, the number of turns on the head will need to be in inverse proportion to the square root of the track width; and, because the flux linking these turns is directly proportional to track width, it follows that the emf from such a head is proportional to the square root of track width. Moreover, for a head of constant inductance, the noise produced by the head amplifier will be constant; and, if it now be assumed that the head amplifier is the only source of noise, then the signal/noise ratio is again proportional to the square root of the width of track.

Since the effect on signal/noise ratio for different track widths is the same whether the noise originates from the tape or the head amplifier, it can be concluded that signal/noise ratio is proportional to the square root of the track width regardless of noise source.

As already mentioned, an error rate of 1 in 10^7 requires a signal/noise ratio of approximately 20 dB.

Conventional 160 μm (0·0063 in) heads provide a signal/noise ratio in excess of 40 dB. Theoretically, therefore, a satisfactory signal/noise ratio should be possible using a track width one hundred times smaller, i.e., 1.6 μm. Though, undoubtedly, edge effects could no longer be disregarded with such a narrow track, this is largely irrelevant; if heads of this width could be produced, which is unlikely, they would tend to wear very rapidly unless braced with a suitable material for spreading the load. Even if this were so there remains the problem of tracking. This might be overcome by recording the tracks using a head of at least double width, so that the tracks are just touching, and by replaying with twin heads slightly staggered. These could be arranged such that if, at any given moment, one head were to be spanning the join between two tracks and hence producing an unuseable signal, the other head would be well clear of the join. However, one fundamental restriction on the spacing of such fine tracks is that, to prevent cross-talk, a head must be separated from adjacent tracks by a significant proportion of the longest wavelength recorded.

In the direction along a track, the size of the head gap must be in proportion to the spacing between one bit and the next, i.e., the digital 'wavelength'. The number of particles influencing the head will be proportional to the square of the bit separation. The square law arises because the magnetic influence penetrates a depth of oxide coating in proportion to wavelength, as well as occupying length. In this case, considering only tape noise, the signal/noise ratio is proportional to the bit spacing.

Consideration of the noise contribution from the head amplifier is very complicated and involves such parameters as the constants associated with the gap, the reluctance of the magnetic circuit, the permeability of the tape, etc. However, it seems reasonable to assume that signal/noise ratio is again linearly related to the bit spacing.

From these conclusions, which have been based on maximising the signal/noise ratio, it is obvious that optimum packing density will be achieved by using extremely narrow tracks, and large bit spacing along the tracks, provided that the tape coating is of adequate thickness. As always, physical limitations will determine what can be achieved in practice.

Practical Experience

The original work was conducted using a recorder having a 2-inch segmented format. Later this was

changed to a 1-inch segmented recorder and attention has recently been turned to a 1-inch non-segmented format. All were analogue machines. It was intended to introduce a digital signal at the input of the record amplifier and, on playback, to recover it from the output of the pre-amplifier, but not to modify the machine in any other way. Obviously it was not possible to re-establish a complete digital picture using an analogue recorder, but this was not essential. A vertical strip of picture was sufficient for a preliminary evaluation.

Initially, one-sixth of the active picture waveform was sampled at three times sub-carrier frequency. Obviously, this could equally have been a quarter of the picture sampled at twice sub-carrier frequency, and was very soon increased to a quarter of the picture sampled at three times sub-carrier frequency. At this time the data was recorded on tape at a rate of 26 Mbit/s, and was coded using delay modulation. An attempt was made to increase the data rate, but eventually the conclusion was reached that the use of rotary transformers for coupling signals to and from the head was incompatible with the low-frequency components which occur with delay modulation. In addition, delay modulation is probably unsuitable since tolerance to timing errors is relatively poor and the eyeheight can be no better than 70%.

A pcm type of signal using nine bits per word was then tried. The incoming 8-bit words were Gray coded, a parity bit was added and following this alternate words were complemented in an attempt to cancel the low-frequency signal. This worked well and the bit rate was increased to 38 Mbit/s giving a capability for $\frac{1}{3}$-width pictures. It was then decided to change the sampling rate to twice sub-carrier frequency which gave more than half the active picture, but certain incompatibilities were revealed between alternate-word complementing and $2f_{sc}$ sampling.

A number of improvements then followed which eventually led to a system which uses the 252 10-bit words having five zeros and five ones, and four other words having six zeros and four ones or vice versa. With this sytem none of the words requires inversion, the dc component is essentially constant, there is an adequate clock component and very little low-frequency energy. There are also, a 100% chance of detecting all single errors, and a 75% chance of detecting all possible errors. The lack of low-frequency components produced by this coding system more than compensates for the higher bit rate required.

All of the work so far described took place using a 2-inch segmented analogue recorder. Use of this served also to prove that halving of the track width was indeed possible; and, so far as could be detected, precisely similar results were obtained. Moreover, it would have been eminently possible to fit twice the number of half-track heads thereby producing a digital recorder capable of dealing with complete pictures, and using the same amount of tape as the analogue version of the equipment, i.e., 2-inch tape at 8 in/s.

During the same period the technique for concealing errors was also improved. The methods adopted were:

i to replace a word in error with the average of the two words one cycle of sub-carrier earlier and later than the error word,

ii to shuffle the words prior to recording so that drop-outs appear as a number of isolated dots lightly spread across the picture should the concealment be switched off,

iii to have a number of start sequences per line so that the loss of a start sequence, or a mis-framing, does not prevent concealment,

iv to provide a technique of fall-back concealment by inserting appropriate parts of a previous line in case drop-outs should occur simultaneously on both tracks of a double-track recorder. The same type of operation is used also for concealing drop-outs of a duration greater than 32 μs in a single-track recorder.

More recently an examination of 1-inch segmented and non-segmented recorders has been made with a view to using them as a basis for a design of digital recorder. This work mostly featured a segmented type of recorder because the speed of rotation of the head wheel could be doubled. With non-segmented recorders it is impossible to do this without the machine becoming a segmented type. In that case, the double-track system would always be appropriate.

Theory versus Practice
Theory indicates that very narrow tracks, with very low bit density per track, yield the optimum packing density. Practice currently shows that relatively broad tracks of 60 μm (0·0024 in) width, with high bit density per track, are satisfactory.

These two statements do not conflict; they merely indicate that a large margin exists for possible future improvement. Whether such improvement will be realised in practice cannot be foreseen. It might be possible only at the expense of head life.

Le Problème de l'Echantillonnage Sub-Nyquist

Résumé

Il paraît probable, à en juger d'après les études intensives dont elles font actuellement l'objet tant en Europe qu'en Amérique du Nord et au Japon, que les techniques numériques finiront par être le moyen prédominant au niveau du studio. C'est pourquoi les membres de l'U.E.R. s'attachent à la définition des normes à adopter pour la télévision numérique. Alors qu'un certain nombre de pays sont en faveur d'un multiplexage des composantes YUV, le Royaume Uni s'est prononcé pour une normalisation basée sur le signal composite PAL dont il prévoit l'utilisation partout où cela est possible avec un échantillonnage à deux fois la fréquence de la sous-porteuse.

Le présent article expose les raisons de ce choix et présente les bases théoriques de la méthode d'échantillonnage sub-nyquist.

Proposition de Normes de Télévision Numériques pour les Signaux PAL en 625 lignes

Résumé

L'emploi de techniques numériques est en train de se répandre rapidement pour gagner tous les stades du traitement et de la transmission du signal vidéo. L'ère du dispositif numérique autonome qui s'insérait dans une chaîne analogique touche à sa fin. Le numérique étant adopté pour des phases de plus en plus nombreuses du traitement, il y a intérêt évident à relier ces phases sans conversion entre les modes numériques et analogiques. D'où la nécessité de connections numériques.

Pour les permettre, les différents paramètres du signal numérique doivent être définis et adoptés non seulement au plan national mais aussi au niveau international et cela aussi rapidement que possible.

Filtres Sub-Nyquist Numériques

Résumé

On propose comme fréquence d'échantillonnage du signal PAL deux normes: quatre fois la fréquence de la sous-porteuse ($4f_{sc}$) sur toute la bande, et deux fois la fréquence de la sous-porteuse ($2f_{sc}$), sub-nyquist. Le présent article décrit les changeurs de fréquence d'échantillonnage numérique d'une mise en oeuvre simple qui permettent de passer de l'une à l'autre de ces deux normes, tout en assurant le filtrage en peigne nécessaire.

Les deux éléments fondamentaux d'un
filtrage en peigne lors de conversions dans les deux sens sont un retard d'une ligne et un filtre passe-bas. On peut faire fonctionner ces deus éléments á $2f_{sc}$ ($8·87$ MHz) dans les deux cas de figure, ce qui permet une économie de matériel considérable. Le retard d'une ligne peut se réaliser sous forme d'un registre à décalage à $8 × 567$eb et le filtrage passe-bas peut être assuré grace à la logique Schottky de faible puissance.

La configuration du filtre passe-bas résulte nécessairement d'un arbitrage entre la performance souhaitée et la complexité du circuit. Celle qui a été choisie aux fins d'expérimentation est du type transversal non-récursif à 13 éléments. Elle comporte environ 75 pacquets de circuits intégrés par filtre, et représente peut-être la plus forte complexité possible dans l'état actuel de la technique.

L'expérimentation doit se poursuivre à l'aide de filtres passe-bas à configuration simplifiée pour en évaluer l'influence sur la qualité de l'image.

Mélangeur Numérique de Télévision

Résumé

Un mélangeur numérique simple a été étudié et réalisé dans le cadre de l'exploration de l'emploi éventuel des signaux PAL échantillonnés à $2f_{sc}$ dans les chaînes Studio.

Dans l'état actuel des connaissances le modèle optimal d'un tel mélangeur utilisera probablement le traitement en série pour la matrice d'affectation en même temps qu'un traitement en parallèle pour le bloc arithmétique. Par ailleurs, à des débits de traitements supérieurs à 80 Meb/s il y a tout avantage à utiliser un dispositif à lignes de transmission. Les signaux de commande de trucage doivent être syncronisés en fonction de l'horloge d'échantillonnage vidéo et un coefficient à 8 bits pour les fondus et le mixage donne une uniformité comparable à celle qu'on obtient avec les méthodes analogiques.

Un filtrage en peigne réglable peut être nécessaire pour la conversion de $2f_{sc}$ à $4f_{sc}$ un retard gênant d'une ligne sur les bords horizontaux des plages de couleur. Les mélangeurs numériques PAL qui assurent le trucage, surtout ceux qui emploient une mémoire à trame, feront vraisemblablement appel à un codage des composantes utilisant la fréquence d'échantillonnage de $4f_{sc}$.

Génération Numérique de Barres de Couleur

Résumé

La nécessité de disposer de signaux d'essais d'origine numérique devient d'autant plus
urgente que l'emploi des techniques numériques se généralise au niveau du studio de télévision. L'élaboration de ces signaux par des moyens numériques est surtout rendue difficile par les caractéristiques des systèmes de télévision en place. Par exemple, dans le système NTSC le signal de chrominance est fourni par une sous-porteuse verrouillée à la fréquence de la ligne qui alterne en phase d'une ligne sur l'autre, alors que dans le PAL, la relation entre la sous-porteuse et la fréquence-ligne est complexe et seule une composante de quadrature du signal de chrominance change de phase d'une ligne sur l'autre. Ces différences de caractéristiques compliquent la tâche de l'ingénieur et peuvent imposer des compromis si l'on doit éviter des solutions couteuses et encombrantes.

Dans le présent article, on expose le problème de la génération numérique de barres de couleur test à des fréquences d'échantillonnage de $4f_{sc}$ et $2f_{sc}$ dans une chaîne PAL. On présente les principes de fonctionnement et les contraintes techniques ainsi que les considérations d'ordre théorique qui influencent la définition correcte des enveloppes de transition. On décrit aussi la méthode fondamentale employée pour insérer le générateur dans une chaîne de démonstration.

Le modèle étudié s'est avéré d'un emploi simple et d'adapte facilement à n'importe quel système PAL dont les caractéristiques sont compatibles; il peut même être utilisé comme source autonome de signaux.

Le problème des Interfaces pour la Distribution Numérique des Signaux Vidéo

Résumé

Les solutions analogiques adoptées pour assurer la distribution de signaux vidéo en studio lancent un défi redoutable aux moyens numériques. Il paraît probable que ceux-ci ne réussiront pas tout-à-fait à la relever et qu'il sera donc nécessaire de rechercher la contre partie des inconvénients économiques en tirant parti des avantages que peuvent offrir à d'autres plans les solutions numériques en studio.

Le matériel de démonstration devait tenir compte de certains besoins particuliers. En gros, on peut envisager trois cas de figure: distribution locale, tous les organes de traitement étant à proximité les uns des autres; distribution à distance, ces organes pouvant être séparés de quelques centaines de mètres, et, la distribution unifilaire, toutes les informations d'interface entre organes étant transmises par un fil unique. Cette dernière solution est la plus satisfaisante, mais peut conduire à une distribution plus coûteuse

que les méthodes analogiques. Les trois cas de figure comportent des besoins différents en formes de signaux, en codage et en caractéristiques de matériel. Ces problèmes sont exposés sommairement en rapport avec les appareils utilisés pour la démonstration organisée par l'U.E.R.

Un Système à faible débit binaire étudié pour la Télévision Numérique

Résumé

Des travaux récents ont mis en évidence les avantages pratiques d'un système de télévision numérique utilisant une fréquence d'échantillonnage Sub-Nyquist (les principes de la méthode d'échantillonnage à $2f_{sc}$ sont exposés ailleurs dans ce numéro). Un tel système permet de ramener cette fréquence à un niveau inférieur à celle qui peut être atteinte pour la transmission sur des chaines numériques à bande limitée: le débit binaire effectif pour une résolution de g^{eb} est d'environ 70 meb/s.

Des signaux sub-nyquist utilisent de façon très efficace la bande vidéo, et sont d'un excellent rendement quand on recherche une bonne qualité avec de faibles débits binaires. Il est également possible de réduire encore davantage de débit pour atteindre 34 meb/s en utilisant un codage différentiel (MICD). Cette valeur a été adoptée comme bonne pour les échanges de programmes internationaux en Europe. Si au Royaume Uni on estime qu'elle n'est pas suffisante pour assurer les transmissions de la plus haute qualité, son utilisation pourrait être envisagée dans ce pays pour les actualités.

Le présent article décrit une méthode permettant d'atteindre ce débit tout en maintenant un bon niveau de qualité.

Le principe mis en œuvre consiste à calculer la valeur de l'échantillon suivant à prélever en fonction d'informations dont on dispose déjà au récepteur. La différence entre la valeur calculée et celle effectivement mesurée est ensuite comprimée, c'est-à-dire le nombre de niveaux quantisés est réduit avant transmission. On dispose donc au récepteur des informations nécessaires pour restituer la valeur primitive de l'échantillon. Du fait du processus de compression une précision accrue de la méthode de calcul conduira à une meilleur qualité de l'image reçue. Par ailleurs, la dégradation de la qualité de l'image affecte surtout les contours qui paraissent "flotter".

Les travaux exposés concernent principalement la précision de la méthode de calcul et l'optimisation de la loi de compression. On explique l'appareillage de calcul simplifié qui a été réalisé en vue de la Démonstration, et suggère des orientations pour les travaux futurs.

La Conversion Analogique-Numérique

Résumé

On dispose essentiellement de deux techniques pour la génération de signaux numériqués échantillonnés à la fréquence sub-nyquist: la première est évidente et consiste à faire fonctionner le dispositif de prélèvement à la fréquence sub-nyquist et à réduire au maximum la dégradation de l'image qui en résulte du fait de "l'aliasing" grâce à un filtrage analogique effectué au niveau de l'interface analogique-numérique. Les filtres utilisés, quoique d'une réalisation qui peut paraître simple, comportent un risque de dérive, ce qui peut accroître plutôt que réduire la dégradation.

L'autre solution consiste à échantillonner le signal vidéo incident à une fréquence supra-nyquist et ensuite à employer un filtrage numérique pour passer au format sub-nyquist recherché. Cette technique permet d'obtenir une précision donnée et une grande stabilité.

Limitation de la Consommation de Bande Magnétique dans un Magnétoscope Numérique

Résumé

On prévoit le remplacement progressif du matériel analogique actuellement utilisé dans les centres de production et sur les réseaux de transmission par des moyens numériques, qui apportent les avantages de la précision, la stabilité et la simplicité. Cette mutation n'était cependant envisageable qu'une fois prouvée la possibilité d'assurer numériquement l'enregistrement des signaux vidéo.

Dans le présent article on compare les caractéristiques essentielles des magnétoscopes analogiques et numériques en s'attachant tout particulièrement aux problèmes que pose le besoin d'obtenir une forte densité d'information sur bande magnétique. Pour les résoudre de manière à disposer d'un enregistreur dont la consommation en bande ne dépasse pas celle d'un magnétoscope hélicoïdal analogique 24mm de qualité radiodiffusion, la théorie veut que le format de la bande soit à pistes très étroites avec une faible densité binaire le long de ces pistes.

Il reste bon nombre de problèmes pratiques à résoudre mais les résultats obtenus jusqu'ici sont prometteurs.

Eine Einführung in die Sub-Nyquist-Abtastmethode

Kurzfassung

Zur Zeit wird in Europa, Nordamerika und Japan die Technologie des Digitalfernsehens praktisch untersucht, und es zeichnet sich ab, dass eines Tages im Studio Digitaltechniken vorherrschen werden. Aus diesem Grunde haben sich die Mitglieder der EBU darauf konzentriert, zukünftige Normen für das Digitalfernsehen festzusetzen. Eine Reihe von Ländern hat ein in Mehrfachschaltung betriebenes YUV-Bauteilformat befürwortet, während Grossbritannien den Wunsch ausgesprochen hat, ein Standardformat zu verwenden, das auf dem komplexen PAL-Fernsehsignal basiert. Grossbritannien hält ausserdem unter Umständen die Verwendung eines PAL-Signal-gemischs für möglich, das mit der Sub-Nyquist-Rate von doppelter Zwischenträgerfrequenz abgetastet wird.

Dieser Artikel untersucht die Gründe für diese Wahl und stellt die theoretische Basis der Sub-Nyquist-Methode vor.

Vorgeschlagene Digitalfernsehnormen für 625-Zeilen PAL-Signale

Kurzfassung

Digitaltechniken dringen rasch in praktisch alle Bereiche der Verarbeitung und Übertragung von Bildsignalen vor. Die Zeit, da unabhängige Digitaleinrichtungen innerhalb einer Analogumgebung verwendet wurden, geht zu Ende. Da jetzt Digitalmethoden in immer mehr Verarbeitungsbereiche eindringen, ist es offensichtlich wünschenswert, diese Bereiche miteinander zu verbinden, ohne die digitalen bzw. analogen Formen umzuwandeln. Die Verbindungen müssen ebenfalls digital sein.

Um dies zu ermöglichen, müssen die verschiedenen Parameter des Digitalsignals definiert, spezifiert und angenommen werden, und zwar nicht nur auf nationaler sondern hoffentlich auch auf internationaler Ebene, und das so frühzeitig wie möglich.

Digitale Sub-Nyquist-Filter

Kurzfassung

Für die Verwendung mit dem PAL-Signal werden zwei Normen vorgeschlagen: Abtasten mit $4f_{sc}$ bei viller Bandbreite und mit $2f_{sc}$ Sub-Nyquist-Frequenz. Im folgenden werden praktikable, digitale Umwandler für diese beiden Normen beschrieben, die gleichzeitig die erforderliche Kammfiltrierung gewährleisten.

Die beiden Grundelemente für die Kammfiltrierung beim Umwandlungsprozess in beiden Richtungen sind eine 1-Zeilen-Verzögerung und ein Tiefpassfilter. Diese beiden Elemente können bei der $2f_{sc}$-Frequenz (8,87 MHz) zur Umwandlung in beiden Richtungen eingesetzt werden, wodurch eine äusserst wirtschaftliche Ausnutzung der Hardware gewährleistet ist. Die 1-Zeilen-Verzögerung kann mittels eines 8×567-Bit Schieberegisters erzielt werden. Die Implementierung des Tiefpassfilters ist unter Verwendung eines Schottky Schwachstrom-Logikschaltkreises möglich.

Der digitale Tiefpassfilter stellt notwendigerweise einen Kompromiss aus der angestrebten Leistung und der Komplexität des Schaltkreises dar. Als Experimentalfilter wählte man nichtrekursive, transversale Modelle mit 13 Abgreifpunkten. Pro Filter resultierte das in einer Zählrate von etwa 75 für den gesamten integrierten Schaltkreis. Das stellt wahrscheinlich die grösste Komplexität dar, die bei Verwendung der heutigen Technologie möglich ist.

Weitere Untersuchungen mit vereinfachten Tiefpassfiltern werden folgen, wobei auch deren Auswirkungen auf die Bildqualität festgestellt werden sollen.

Ein Mischer für Digitalfernsehen

Kurzfassung

Als Tiel einer Untersuchung der potentiellen Verwendung von mit $2f_{sc}$ abgetasteten PAL-Signalen in Studio-Systemen ist ein einfacher Digitalmischer entworfen und gebaut worden.

Die optimale Konstruktion eines solchen Mischers dürfte zum gegenwärtigen Zeitpunkt Serienverarbeitung für die Zuordnungsmatrix, jedoch Parallelverarbeitung für die Recheneinheit einschliessen. Überdies können bei Verarbeitungsraten von über 80 MBit/s durch die Verwendung von Übertragungsleitungsanordnungen beträchtliche Vorteile erzielt werden. Die Kontrollsignale für besondere Effekte sollten in Beziehung gesetzt werden zu einem Videoabtastzeitgeber. Ein 8-Bit-Koeffizient zum Regeln und Mischen ergibt eine Konstanz, die mit Analogverfahren vergleichbar ist.

Bei der Umwandlung von $2f_{sc}$ nach $4f_{sc}$ ist unter Umständen adaptive Kammfiltrierung erforderlich, um unerwünschte 1-Zeilen-Farbtonverzögerungen an den horizontalen Tasträndern zu vermeiden. PAL-Digitalmischer für komplexe Sondereffekte, insbesondere die mit einem Bildspeicher, werden eher ein auf $4f_{sc}$ ausgerichtetes Teilkodiersystem verwenden.

Digitale Farbbalkenerzeugung

Kurzfassung

Mit der ständig zunehmenden Verwendung von digitalen Methoden in Fernsehstudio-Systemen nimmt auch der Bedarf an digital erzeugten Testsignalen zu. Schwierigkeiten hierbei beruhen zum grossen Teil auf dem jeweiligen Fernsehsystem, das zugrundegelegt wird. Beim NTSC-System wird das Buntsignal beispielsweise von einem Zwischenträger geliefert, der zeilenabhängig ist und von Zeile zu Zeile in der Phase wechselt. Beim PAL-System hingegen ist die Beziehung zwischen Zwischenträger und Zeilenfrequenz komplex, und nur eine Quadraturkomponente des Buntsignals erfährt eine Phasenverschiebung von Zeile zu Zeile. Derartige Eigenschaften bereiten dem Konstrukteur Schwierigkeiten und können Kompromisslösungen erforderlich machen, wenn man Unhandlichkeit und hohe Kosten vermeiden will.

In diesem Artikel werden die Probleme der Erzeugung von digitalen Farbbalken-Testsignalen mit Abtastfrequenzen von $4f_{sc}$ und $2f_{sc}$ zur Verwendung in einer PAL-System-I-Umgebung erörtert. Grundlegende Betriebsprinzipien, Konstruktionsgrenzen und die theoretischen Überlegungen für die Erzielung zufriedenstellender Übergangshüllkurven werden ebenso erörtert wie die zur Intergrierung des Generators in ein Vorführsystem verwendete prinzipielle Methode.

Die Konstruktion beweis Flexibilität im Einsatz und erwies sich als leicht anpassungsfähig an jedes digitale PAL-System mit kompatiblen Eigenschaften, selbst als in sich abgeschlossene Signalquelle.

Grenzebenen bei Digitaler Video-Verteilung

Kurzfassung

Analogverfahren zur Verteilung von Videosignalen im Studio haben Digitalverfahren einiges voraus. Die Digitalverfahren werden diese Kluft wahrscheinlich nicht schliessen können. Deshalb muss aus wirtschaftlichen Gründen bei anderen Einrichtungen des Digitalstudios ein Ausgleich gefunden werden.

Bei der Vorführausrüstung mussten einige besondere Anforderungen in Betracht gezogen werden. Allgemeingesprochen kann man die Probleme der Studioverteilung von drei Standpunkten aus betrachten. Erstens Nahverteilung, wo die wichtigsten Verarbeitungssysteme eng beieinander liegen; zweitens Fernverteilung, wo diese Systeme bis zu einigen Hundert Metern auseinander liegen; dritten Verteilung über eine Einzelleitung, wo alle Grenzebeneninformationen zwischen den Systemen über eine Einzelleitung laufen. Letzteres ist die beste Lösung. Sie kann jedoch im Vergleich zu Analogmethoden zu höheren Verteilungskosten

führen. Alle drei Möglichkeiten stellen verschiedene Anforderungen in bezug auf Signalformate, Kodieranordnungen und Hardware. Diese Gesichtspunkte werden im Zusammenhang mit der für die EBU-Vorführung verwendeten Ausrüstung kurz erörtert.

Ein System mit niedrigen Bit-Mengensätzen für Digitalvideo

Kurzfassung

Jüngste Arbeiten haben die Brauchbarkeit eines digitalen Fernsehsystems bewiesen, das mit einer Sub-Nyquist-Abtastfrequenz arbeitet. Die Prinzipien dieses $2f_{sc}$-Abtastverfahrens werden an anderer Stelle in dieser Ausgabe beschrieben. Ein derartiges System reduziert die Abtastrate auf ein Mass, das niedriger ist, als es sonst für die Übertragung auf bandbegrenzten Digitalsystemen möglich wäre. Die effektive Datenfrequenz bei 8-Bit-Auflösung beträgt ungefähr 70 MBit/s.

Sub-Nyquist-Signale nutzen das Video-Spektrum sehr gut aus, und vieles spricht für sie, wenn es um hohe Qualität bei niedrigen Bit-Raten geht. Es besteht ausserdem die Möglichkeit, die Datenfrequenz weiter bis auf 34 MBit/s zu senken, indem man differentiale impulsverschlüsselte Modulation verwendet. Diese Datenfrequenz ist als Norm für den internationalen Programmaustausch innerhalb Europas angenommen worden. Obschon man sie in Grossbritannien für Übertragungen bester Qualität als unzureichend erachtet, könnte man sie hier bei der Nachrichtensammlung einsetzen. Dieser Artikel beschreibt ein Verfahren, mit dem bei gleichzeitiger Aufrechterhaltung eines hohen Qualitätsniveaus diese Bit-Frequenz erzielt werden kann. Folgendes Prinzip wird dabei angewendet: der Wert des jeweils nächsten Abtastpunktes wird vorausgesagt, wobei man von Informationen ausgeht, die bereits am Empfänger erhältlich sind. Die Differenz zwischen dem vorhergesagten Wert und dem tatsächlichen Wert des Abtastpunktes wird dann komprimiert, d.h. die Anzahl der gequantelten Stufen wird vor der Übertragung reduziert. Auf diese Weise sind am Empfänger genügend Informationen verfügbar, um den ursprünglichen Wert des Abtastpunktes zu konstruieren. Aufgrund des Komprimierungsprozesses hat grössere Genauigkeit der Voraussagemethode eine bessere Bildempfangsqualität zur Folge. Die verminderte Bildqualität steht darüberhinaus hauptsächlich im Zusammenhang mit Konturen im Bild und äussert sich als sogenanntes Randflimmern.

Die hier beschriebene Arbeit beschäftigt sich hauptsächlich mit der Genauigkeit der Voraussagemethode und der Optimierung des Komprimierungsgesetzes. Darüberhinaus wird der für die Vorführung entwickelte einfache Prädiktor erklärt, und es werden Anregungen gegeben.

Analog-Digital-Umwandlung

Kurzfassung

Für die Erzeugung von mit der Sub-Nyquist-Rate abgetasteten Digitalsignalen gibt es zwei Grundverfahren. Das erste Verfahren liegt auf der Hand: hierbei arbeitet man mit der Sub-Nyquist-Abtastfrequenz und setzt die sich aufgrund des Aliaseffektes ergebende Bilddegradation auf ein Mindestmass herab, indem man an der analog-digitalen Grenzebene analoge Filterverfahren anwendet. Derartige Filter scheinen zwar einfach zu konstruieren zu sein, können jedoch zu Abweichungen neigen, was eher zu einer stärkeren als zu einer geringeren Degradation führt.

Beim zweiten Verfahren tastet man die eintreffenden Video-Signale oberhalb der Nyquist-Rate ab. Anschliessend verwendet man digitale Verfahren zur Filterung und Umwandlung in das erforderliche Sub-Nyquist-Format. Auf diese Weise werden definierbare Präzision und hohe Stabilität sichergestellt.

Geringer Bandverbrauch für Digital-VTR

Kurzfassung

Man geht davon aus, dass die zur Zeit in Studios und zu Übertragungszwecken verwendeten analogen Systeme schrittweise durch digitale Systeme ersetzt werden, die bestimmte Vorteile wie Präzision, konsequente Leistung und einfachen Betrieb bieten. Dieser Wandel ist jedoch erst vorstellbar geworden, als sich die digitale Video-Aufzeichnung als durchführbar erwies.

In diesem Artikel wird ein Vergleich gezogen zwischen den wesentlichen Merkmalen analoger und digitaler Rekorder, mit dem Schwergewicht auf den Problemen und Erfordernissen, die entstehen, wenn man hohe Informationsdichten auf Magnetband erzielt. Bei der Überlegung, wie diese Fragen beantwortet werden können, damit man einen Digitalcomputer erhält, der nicht mehr Band verbraucht als ein gegenwärtig zum Senden verwendeter Schnecken-scan-Analogrekorder, ist theoretisch ein Bandformat mit sehr engen Spuren erforderlich, das jedoch auf diesen Spuren eine geringe Bit-Dichte aufweist.

Viele praktische Probleme müssen noch bewältigt werden; die bisherigen Ergebnisse sind jedoch vielversprechend.

Introducción al Muestreo Sub-Nyquist

Resumen

Considerando que la tecnología televisiva digital es actualmente objeto de activo análisis ponderado en Europa, Norteamérica y Japón, todo apunta a un predominio futuro de las técnicas digitales en los estudios de TV y por ello los miembros de la EBU han venido concentrándose en la evaluación de la normalización previsible para transmisiones de este tipo. En tanto que diversos países se inclinan por un formato multiplexado de componentes YUV el Reino Unido ha expresado su preferencia por una norma basada en el PAL compuesto y anticipa, en todos los casos aplicables, el uso de una señal PAL compuesta muestreada en régimen sub-Nyquist al doble de la frecuencia subportadora.

En este artículo se examinan las razones en que se fundamenta tal elección y se expone la base teórica del método de muestreo sub-Nyquist.

Normas Digitales de TV Propuestas para señales PAL de 625 líneas

Resumen

Las técnicas digitales están invadiendo con celeridad prácticamente todos los sectores de transmisión y proceso de señales generadoras de imágenes. La era de los dispositivos digitales aislados que funcionan entre equipos analógicos está llegando a su término. Al compás que penetran en más y más dominios del proceso de señales los métodos digitales se acentúa la conveniencia de interconectar estos campos sin necesidad de interconversión digital-analógica. Tales interconexiones han de ser, asimismo, digitales.

Si queremos plasmar este objetivo los diversos parámetros de la señal digital han de ser objeto de definición, especificación y acuerdo, a escala no ya nacional sino esperemos que también internacional, a la mayor brevedad posible.

Filtros Digitales Sub-Nyquist

Resumen

El muestreo en régimen cuádruple de frecuencia subportadora en todo el ancho de banda y doble de frecuencia subportadora a tenor del criterio sub-Nyquist son dos normas propuestas de aplicación a la señal PAL. El texto describe unos interconversores digitales de régimen de muestreo muy prácticos para pasar de una a otra norma y proporcionar al mismo tiempo el filtraje de peine necesario.

Los dos elementos esenciales para llevar a cabo el filtraje de peine en conversión ascendente y descendente son un retardo de una línea y un filtro de paso bajo; es posible la actuación de ambos en régimen doble de la frecuencia subportadora (8,87 MHz) para la conversión directa o inversa, con lo que se logra un ahorro considerable en el capítulo de equipo. El retardo monolineal es susceptible de materialización en forma de un registrador de desplazamiento de 8×567 bitios y para el filtro de paso bajo se recurre a lógica de Schottky en bajo desarrollo.

El esquema del filtro de paso bajo digital responde necesariamente a un punto medio entre el rendimiento deseable y la complejidad del circuito. Para los filtros experimentales descritos se eligió un esquema transversal no recursivo de 13 tomas, que arrojó una cuentabloque de circuitos integrados en el entorno de los 75 por filtro y tal vez sea exponente del mayor grado de complejidad factible dentro de nuestra actual tecnología.

Se realizarán más trabajos sobre esquemas simplificados del filtro de paso bajo y la evaluación de su efecto sobre la calidad de la imagen.

Mezclador Digital de TV

Resumen

Dentro del programa de investigación del uso en potencia de señales PAL muestreadas en régimen $2f_{sp}$ se ha proyectado y construido un mezclador digital de sencilla concepción.

Actualmente el esquema óptimo de un mezclador de esta índole tiende a recurrir al proceso en serie para la matriz de asignación y en paralelo para la unidad aritmética. Por añadidura, tenemos que para velocidades de proceso superiores a los 80 Mbitios/seg. se obtienen sensibles ventajas utilizando técnicas de distribución de línea de transmisión. Las señales de control para los efectos especiales han de estar gobernadas por el reloj vídeomuestreado y un coeficiente de 8 bitios para mezcla y desvanecimiento es fuente de uniformidad comparable a la conseguida mediante técnicas analógicas.

Si se desea evitar la molesta presencia de un retardo monolineal de saturación cromática en los frentes horizontales de manipulación conviene considerar la incorporación del filtraje de peine adaptador de $2f_{sp}$ a $4f_{sp}$. Los mezcladores digitales PAL que dan complejos efectos especiales, y en particular los que operan con memoria de cuadros, suelen emplear un sistema codificador de componentes orientado al valor cuádruple de la frecuencia subportadora.

Generacion de Barra de Color Digital

Resumen

El uso cada vez más difundido de las técnicas digitales en los estudios de TV viene haciendo más apremiante la necesidad de señales de prueba dígitogeneradas y la dificultad de obtenerlas es función, mayormente, del sistema televisivo en cuestión. Por ejemplo, en el NTSC la señal de crominancia la provee una subportadora que está bloqueada en línea y que alterna en fase de una línea a otra, mientras que en el PAL la subportadora y la frecuencia de línea tienen una relación compleja y tan sólo un componente en cuadratura de la señal de crominancia cambia de fase de línea a línea. Las características de este tipo representan complicaciones para el ingeniero de proyecto y pueden conducir a soluciones de compromiso para evitar tamaños desmesurados y costes excesivos.

En este artículo se debaten los problemas inherentes a la generación digital de señales de prueba de barra de color con frecuencias de muestreo de $4f_{sc}$ y $2f_{sc}$ para su utilización en el ámbito del PAL System I. Asimismo, se analizan los principios operativos básicos, las limitaciones en el trazado del esquema y los conceptos teóricos para establecer envolventes de transición satisfactorias, al par que el método esencial de incorporar el generador a un sistema de demostración.

En la práctica el sistema demostró su flexibilidad operativa y evidenció la posibilidad de fácil adaptación a cualquier sistema PAL digital de características compatibles o incluso al funcionamiento como fuente autónoma de señales.

Interfases para Videodistribución Digital

Resumen

Las técnicas analógicas de distribución de vídeoseñales en los estudios presentan un considerable desafío a sus equivalentes digitales; parece que estas últimas no podrán superarlo plenamente y desde un prisma de racionalización económica el equilibrio habrá de restablecerse en otros sectores digitales del estudio.

. En cuanto al equipo de demostración hubo que calibrar ciertos planteamiento especiales. Las dificultades de distribución en el estudio pueden esbozarse desde tres enfoques: El de la distribución local, en que los grandes equipos de proceso se hallan próximos entre sí; el de la distribución a distancia, en que su separación puede ser de varios cientos de metros, y el de la distribución monofilar en la que toda la información de interfase circula por un solo cable de enlace entre unidades. Este esquema

es el más conveniente pero puede originar costes de distribución mayores que los que implican los métodos analógicos. Cada uno de estos tres enfoques entraña requisitos distintos en lo referente a los formatos de señal, procedimientos de codificación y el equipo, los cuales se examinan brevemente en el contexto de las máquinas utilizadas para la demostración EBU.

Transmisión lenta de bitios para Vídeo Digital

Resumen

Trabajos de reciente realización han demostrado el carácter práctico de un sistema digital de TV que opere con frecuencia de muestreo sub-Nyquist y en otras páginas de este número se exponen los principios del método de muestreo a $2f_{sp}$. El sistema aludido reduce el régimen de muestreo a un nivel menor del necesario para efectuar la transmisión mediante sistemas digitales con limitación de banda; la velocidad efectiva de circulación de los datos para una resolución de 8 bitios es de unos 70 Mbitios/seg.

Con señales sub-Nyquist se obtiene un magnífico rendimiento del vídeoespectro y excelentes perspectivas de óptima calidad a bajos regímenes de transmisión. Existe, igualmente, la posibilidad de reducir todavía más el régimen a 34 Mbitios/seg. con modulación diferencial de código de impulsos (mdci), velocidad adoptada como norma para el intercambio internacional de programas dentro de Europa. Aunque en el Reino Unido esta velocidad no se considera que resulte adecuada para transmisiones de óptima calidad podría utilizarse aquí para recoger noticias. En este artículo se describe un método de conseguir un elevado nivel de calidad a este régimen, sobre el principio de pronosticar el valor de la muestra siguiente que se va a tomar en base a la información ya disponible en el receptor. La diferencia entre el valor pronosticado y el real de la muestra se somete seguidamente a compresión, esto es, se reduce el número de niveles cuantificados antes de transmitir. Así, se tienen en el receptor datos suficientes para reconstruir el valor original de la muestra. Gracias al proceso compresor una mayor exactitud del método de pronóstico redunda en imágenes de mejor calidad en el receptor y, además, la pérdida de calidad en la imagen recibida es en esencia, inherente a los contornos, que pueden aparecer "recargados".

El trabajo aquí descrito se centra principalmente en la precisión del método de pronóstico y la optimización del proceso compresor, trata del sencillo aparato pronosticador construido para la demostración y formula algunas sugerencias dignas de tenerse en cuenta a efectos de consideración más detallada.

Conversión Analógico-digital

Resumen

Para la generación de señales digitales muestreadas en régimen sub-Nyquist existen dos técnicas fundamentales. La primera es el método obvio de situar el dispositivo muestreador en régimen de funcionamiento sub-Nyquist y reducir al mínimo la consiguiente degradación de la imagen debida al efecto de copia (aliasing) mediante técnicas de filtraje analógicas en la interfase analógico-digital. Estos filtros, aunque sencillos de concepto en apariencia, son susceptibles de desvío que incremente la degradación en vez de reducirla.

La alternativa es muestrear la vídeoseñal de entrada por encima del régimen Nyquist y recurrir a técnicas digitales para filtrarla y convertirla al formato sub-Nyquist requerido, garantizando de esta forma una elevada estabilidad y precisión definible.

Aprovechamiento de Cinta en Grabaciones Vídeo

Resumen

Todo hace prever el progresivo reemplazo de los equipos analógicos actualmente utilizados en los centros televisivos y la transmisión intercadena por equipos digitales que se distinguen por su precisión, rendimiento uniforme y sencillez de manejo. Pero tal cambio no podía llegar hasta que se pusiera de manifiesto la viabilidad de las vídeograbaciones digitales.

Este artículo aborda la comparación entre las características esenciales de los registradores digitales y analógicos y, en especial, las dificultades y requisitos que se derivan del registro superdenso en cinta magnética. En la consideración de procedimientos para resolver dichas dificultades, a fin de obtener una máquina digital que no use más cinta que uno de los actuales registradores analógicos, de escansión helicoidal, de emisión de 1 pulg. (25,4 mm.), las directrices teóricas apuntan a un formato de cinta con canales muy estrechos y baja densidad de bitios a lo largo de ellos.

Aunque todavía resten muchos problemas prácticos por solventar los resultados conseguidos hasta la fecha son alentadores.

IBA TECHNICAL REVIEW

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